

TM9-5000-9

DEPARTMENT OF THE ARMY TECHNICAL MANUAL

NIKE I SYSTEMS ACQUISITION RADAR CIRCUITRY (U)



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[AG 413.44 (3 Apr 56)]

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NG: None.

USAR: None.

For explanation of abbreviations used, see SR 320-50-1.

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CHAPTER 1.

INTRODUCTION

Section I. DESCRIPTION OF THE SYSTEM

1. PURPOSE

This text contains instructional material on the functioning and circuitry of the acquisition radar. This material has been compiled with the intent of forming an adequate basis of instruction for maintenance personnel whose duties will require them to keep the acquisition radar in working order.

2. SCOPE

This text covers the functional and control circuitry of the acquisition radar in block level detail and in functional circuit detail of the individual components which make up the various systems into which it is subdivided. In addition to functional circuit explanation, field adjustments of individual chassis are given.

3. REFERENCES

The principal source of material for this text is in school lesson plans of the 5600 series, modified by material gleaned from the Bell Telephone Laboratories (BTL) texts. System and circuit references given in the text refer to schematic volumes. TM 9-5000-26 is the principal schematic reference, and references to it are made by page number. Other schematic references (to TM 9-5000-25) are made by running sheet number, referred to as sheet number. References to figure refer to figures which are a part of this text.

4. HISTORY

In the closing days of World War II there appeared in the skies of western Germany the first of a new class of aircraft. These were extremely fast (although very short lived) for that day. They flew by jet power. The appearance of these aircraft was the forerunner of new requirements to be met in the solution of the antiaircraft problem. Now, it has become possible for modern jet-propelled aircraft to approach a target area at or near the speed of sound and release their bombs while yet out of range for projectile-firing defense guns. It is perfectly feasible for the same aircraft to maneuver at long ranges and high altitudes, with

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respect to defending guns, in a manner such as to render accurate intercept point prediction for the projectile impossible. The missile used to combat these supersonic, high-flying jet aircraft must be capable of changing course during flight to cope with the new type target, and in addition it must have the advantage of its intended target in speed, maneuverability, and operating altitude. These specifications have led to the development of the Nike I system as a solution to the problem. To meet the requirements set up for solution of the modern anti-aircraft problem, the Nike I fire control system has been developed. In this text, the reader will be concerned primarily with only the integrated fire control (IFC) equipment. There are three radars used in the battery control area, one of which is the acquisition radar (fig 1), a search-type radar operating the S-band (3,100 to 3,500 megacycles per second). The acquisition radar has a pulse repetition frequency of 1,000 pulses per second and a peak r-f power output of 1 megawatt. It scans the entire horizon for ranges up to 120,000 yards to engage targets and, by means of the Mark X IFF equipment included with it, establishes their identity as friend or foe. By means of electronic circuitry and equipment, to be explained later in this text, a target identified as foe can be transferred to a target-tracking radar that will track it for engagement by a Nike missile which will remove it from the sky.

5. DESCRIPTION

The acquisition radar is composed of units which are located in the battery control trailer and others which are suspended beneath the antenna drive unit in the antenna tripod assembly which comprises the acquisition barrette. Units located in the battery control console and the acquisition power control panel control the radar. Power and control circuitry are carried to the barrette on five cables which connect the battery control trailer to the acquisition barrette. The reflected radar signals are displayed on two PPI-type, 10-inch tubes and two B-type precision indicators. One of each is located on the battery control console; the other two are located in the radar control trailer on the target console. This presentation appears in both locations along with an electronically generated range circle and radial azimuth line. Both of these electronic markers are steerable from the battery control console. The range circle can be set for a range of either 60,000 yards or 120,000 yards. The azimuth marker can be steered around the full 360° of the face of the indicator. The battery control console personnel use these markers to designate rapidly any target visible on the indicators to the operators at the target console. These and other features of the acquisition radar will be explained in six sections:

- Synchronizing System.
- Transmitter System.
- Acquisition Antenna System.
- Acquisition Receiver System.
- Presentation System.
- Target Designation.

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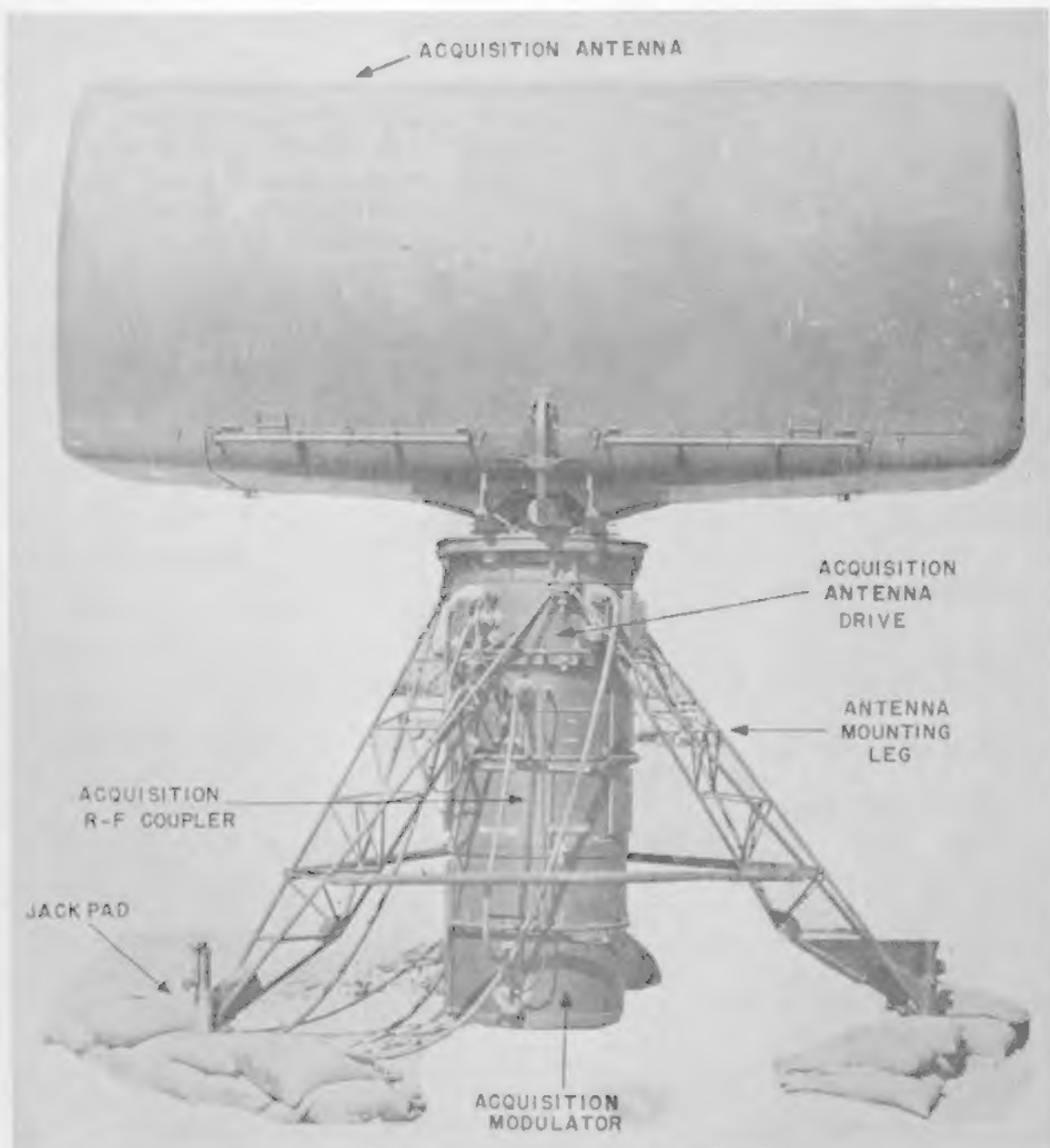


Figure 1. Acquisition antenna assembly.

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6. ACQUISITION RADAR CHARACTERISTICS

ANTENNA.

Type	Pillbox-reflector type.
Beam	Two vertical beam shapes obtainable: (1) A long pencil-shaped pattern having a verticle beam width of 97.8 mils at the one-way, half-power points is available to 284 mils of elevation, and (2) a cosecant-squared pattern covering a 713-mil sector is available at 157 mils of elevation.
Horizontal beam width	25-mil width at the one-way, half-power points.
Gain	35 decibels.
Polarization	Horizontal.
Horizontal pump system:	
Pressure	300 psi.
Rated voltage	208v, three-phase, 400-cycle.
Limits of operation:	
Azimuth (both beams)	Continuous operation. Rotational speeds of 10, 20, or 30 rpm.
Elevation:	
Pencil beam	36 mils to 249 mils.
Cosecant-squared beam	107 mils to 392 mils.
Range:	
Pencil beam	140,000 yards (target of B47 size).
Cosecant-squared beam	100,000 yards (target of B47 size).
Rates (Remote Range Data).	
Tracking	1,000 yd/sec.
Slewing	12,000 yd/sec.

TRANSMITTER SYSTEM.

Frequency	S-band, 3, 100 to 3, 500 mc, tunable over 12 percent band.
Transmitter	Tunable magnetron.
Peak r-f power	1 megawatt.
Pulse repetition frequency	1,000 pulses per second.
Pulse duration	1.3 microseconds.
Modulator type	Hydrogen thyatron, line-type.
R-F lines:	
Width	3 inches.
Height	1 1/2 inches.

RECEIVER SYSTEM.

Antenna coupling	TR and ATR duplexing assembly.
Receiver type	Double detection.
First detector	IN28 rectifying crystal unit.
Local oscillator type	6BL6 reflex oscillator with plug-tuned external cavity.
I-F frequency	60 megacycles.
I-F bandwidth	2 mc at 3 db down.
Over-all receiver noise figure	15 decibels.
Conversion loss	8 decibels.

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Section II. CIRCUITRY COMPONENTS

7. SYNCHRONIZING SYSTEM

The synchronizer unit is located in the battery control trailer in the acquisition radar cabinet assembly. The physical location of the synchronizer is shown on page 191, TM 9-5000-26, and its functional place in the radar, on page 188. A detailed schematic of the synchronizer is shown on page 198 of the same special text. The synchronizer delivers a 40-volt, 2-microsecond, positive preknock pulse occurring at the rate of approximately 1,000 pulses per second. The preknock pulse is used for several purposes throughout the acquisition and target-tracking radars; the main purpose of preknock is to initiate the sweeps on the PPI and the precision and tracking indicators. In addition to the preknock pulse, the synchronizer delivers a positive, 1.5-microsecond, 40-volt sync pulse also occurring at the rate of approximately 1,000 pulses per second. This pulse is delayed 23.5 microseconds after the preknock pulse. Thus, in the sequence of events, each preknock pulse will be followed after 23.5 microseconds by a sync pulse. The sync pulse is used to determine the time at which the transmitters (acquisition, target-tracking, and IFF) will fire. The 23.5-microsecond delay from preknock to sync pulse permits the sweeps on the indicators to be established by the time the transmitters are fired.

8. TRANSMITTER SYSTEM

The components of the transmitter system are located in the two lower cans of the acquisition barrette, and in the acquisition radar cabinet assembly of the battery control trailer. In TM 9-5000-26 a block diagram of the transmitter system is found on page 195. The function of the acquisition transmitter is to produce high power pulses of r-f energy. The r-f system radiates these pulses into space in a definite pattern. The pulses contain approximately 5,000 kw of energy and are 1.3 microseconds in duration, with a prf of 1,000/sec at the time of the pulse. It is the reflection of the pulses of r-f energy, amplified by the receiving system, which allows the acquisition radar to detect and display targets on the acquisition presentation system. The transmitter system consists of a magnetron and the channels necessary to trigger it. In order to develop a pulse of sufficient amplitude, the transmitter employs special types of circuits. The components which make up the transmitter system are the trigger generator, modulator, high-voltage power supply, pulse transformer, and magnetron, with the magnetron tuning drive used for tuning the magnetron to the desired (or assigned) frequency. The r-f system is shared between the transmitter and receiver system. It is composed of the following: two ATR tubes, one TR tube, a 10-cm waveguide, and a cosecant bar injection-type antenna. These components convey the transmitted energy from the magnetron into space.

9. ACQUISITION ANTENNA SYSTEM

The acquisition antenna is provided to beam the pulsed r-f energy into space. Details of this assembly are shown in TM 9-5000-26, page 190, and other pertinent references may be found in the same text on page 187, location index. The antenna provides continuous surveillance of the assigned defense area. It is rotated in azimuth at speeds of 10, 20, or 30 rpm, and is simultaneously tilted at a rate of approximately 1 1/2 cycles per minute over a range of 20 degrees, +2 to +22 degrees above the horizon (figs 14-17). Within the antenna the magnetron operates through a system of waveguides and out into space. The radome contains a parabolic cylindrical reflector having a surface made of two sets of horizontal reflecting bars. The main bars are fixed to form the parabolic cylindrical surface which creates a long-range pencil-shaped beam pattern providing coverage out to the range of 120,000 yards. The second set of horizontal reflecting bars is moved between the fixed bars in the lower section of the antenna to reshape the surface and provide a cosecant-squared pattern furnishing high altitude coverage of 60,000 feet at short range. The antenna functions in a dual manner, transmitting the r-f pulses and receiving the r-f echoes. The two ATR tubes and the TR tube in the duplexer assembly provide protection to the receiver from the transmitted pulses and prevent the return echo from being dissipated in the transmitter. Four elevation scan modes are provided to vary the elevation coverage (figs 14-17). The pencil beam switches over to a cosecant-squared pattern at a different elevation angle for each mode. This switchover is brought about by the injection of the movable set of horizontal reflecting bars. Each mode has a different value of maximum beam elevation.

10. RECEIVER SYSTEM

A block diagram of the receiver system may be found in TM 9-5000-26, page 207. The purpose of the receiver system is two-fold, to amplify the weak-strength return echos to a usable level and provide means to change the r-f echo pulses into video that may be presented on the indicators. Inasmuch as no practicable means has been devised to amplify the ultrahigh frequencies involved (3,100-3,500 mc), the receiver used is of a conventional superheterodyne type. This superheterodyne receiver changes the frequency of the return echo to lower frequency (60 mc) and then amplifies that lower frequency. The receiver then detects the pulse envelope, further amplifies these pulses, and displays them on the PPI and PI.

11. MOVING TARGET INDICATOR SYSTEM (MTI)

The purpose of the moving target indicator is to eliminate fixed target echoes from the acquisition radar indicators, thus permitting the operator to detect moving targets in high-clutter areas. The range to which the stationary targets will be eliminated and normal operation resumed can be controlled by the operator (0 to 35,000 yards). All of the components of the MTI system are located in the battery control trailer in the acquisition radar cabinet assembly unit. This system consists of the delay cell amplifier, the nondelay amplifier, MTI synchronizer, delay cell, and delay amplifier. Schematics can be found by referring to the receiver index, page 206 in TM 9-5000-26. The MTI system makes use of the fact that the amplitude of the echo received from fixed targets remains relatively constant from pulse to pulse, while the echo amplitude of a moving target varies. Each received echo is added algebraically to its corresponding preceding inverted echo. If the echo amplitudes of the two successive echoes are equal, their algebraic sum is zero, and nothing appears on the scopes. However, if the echo amplitudes of the two successive echoes are unequal, their algebraic sum is not zero, and this amplified sum appears on the scopes.

12. RANGE SYSTEM

The range of a target detected by the acquisition radar is determined by measuring the time interval between the transmission of the pulse (main bang) and the reception of the echo from the target in question. Extreme accuracy is unnecessary for this time measurement, since acquisition range data are used only for target designation and other surveillance functions. The accuracy required for the acquisition range determination system is ± 150 yards, corresponding to a time measurement accuracy of ± 0.9 microsecond. The time measurement is accomplished by aligning a continuously movable range marker, the range circle, with the target pip. This range mark, which is a sharp pulse occurring after each main bang, is generated in the acquisition range unit; the time delay between the main bang and the range mark is controlled by a potentiometer geared to a dial calibrated in yards. The range potentiometer is installed in the target designate control panel, which is located in the battery control drawer. See page 223, TM 9-5000-26, for the schematic of the control panel. The potentiometer is geared directly to the handwheel on the control drawer marked RANGE and to the acquisition range dial on the control drawer. The acquisition range unit is connected electrically to the range potentiometer in the acquisition range control unit. The circuits of the acquisition range unit are triggered by the preknock pulse, generated by the synchronizer 23.5 microseconds before the transmitted or sync pulse (main bang). The unit yields two outputs: the acquisition range mark, a sharp, positive pulse whose delay with respect to the input trigger is controlled by the range potentiometer and may vary between about 44 microseconds

and 756 microseconds, corresponding to ranges between 3,500 yards and 120,000 yards; and the acquisition range gate, an 8,000-yard positive pulse, the leading slope of which precedes the acquisition range mark by 24 microseconds. The acquisition range mark is supplied to the PPI's in the battery control trailer and the radar control trailer, where it appears as the range circle; it is also supplied to the precision indicator B-scope in the battery control trailer where it is displayed as the horizontal acquisition range line. Finally, the range mark is sent to the target-tracking ranging system for target transfer purposes. The acquisition range gate is also supplied to the precision indicator in the battery control trailer where it is used for generating the expanded range sweep.

13. PRESENTATION SYSTEM

a. Acquiring target data. Antiaircraft gun batteries have in the past been equipped with radar and fire control devices which have in many respects been unsatisfactory. One great disadvantage (with the exception of the T33 and M33 systems) has been the inability of earlier systems to detect the approach of one target while actually engaging another. In the last war, many instances occurred in which a battery, while unsuccessfully engaging a target near maximum range, discovered too late the presence of a nearer target against which its fire would have been more effective. The Nike I acquisition radar is designed to provide the battery control officer with data concerning all targets near the battery and to provide these data even while the battery is engaging a target. Thus, the fire of the battery may always be directed upon the target which can be engaged with the greatest chance of success or upon the target which constitutes the greatest threat to the defended area. These target data are displayed on the indicators of the acquisition presentation system (one PPI and one B-scope precision indicator in the battery control trailer and one of each in the radar control trailer); but by itself, it is not enough. A means must be provided by which the range and azimuth of any target may be determined and by which the range and azimuth of a target selected for engagement may be designated to the target-tracking radar. These are the functions of the range circle and steerable azimuth line. In addition, the displayed data must indicate the position of the target being tracked by the target-tracking radar. This position is indicated by the electronic cross, which appears on the PPI display at the range and azimuth corresponding to the range and azimuth setting of the target-tracking radar. These elements of data, displayed on a PPI, are of great value in enabling the battery control officer to make competent decisions. To present a more detailed view of a particular sector appearing on the PPI display, an expanded B-scope presentation is provided by two 5-inch precision indicators, presentation system block, page 221, TM 9-5000-26.

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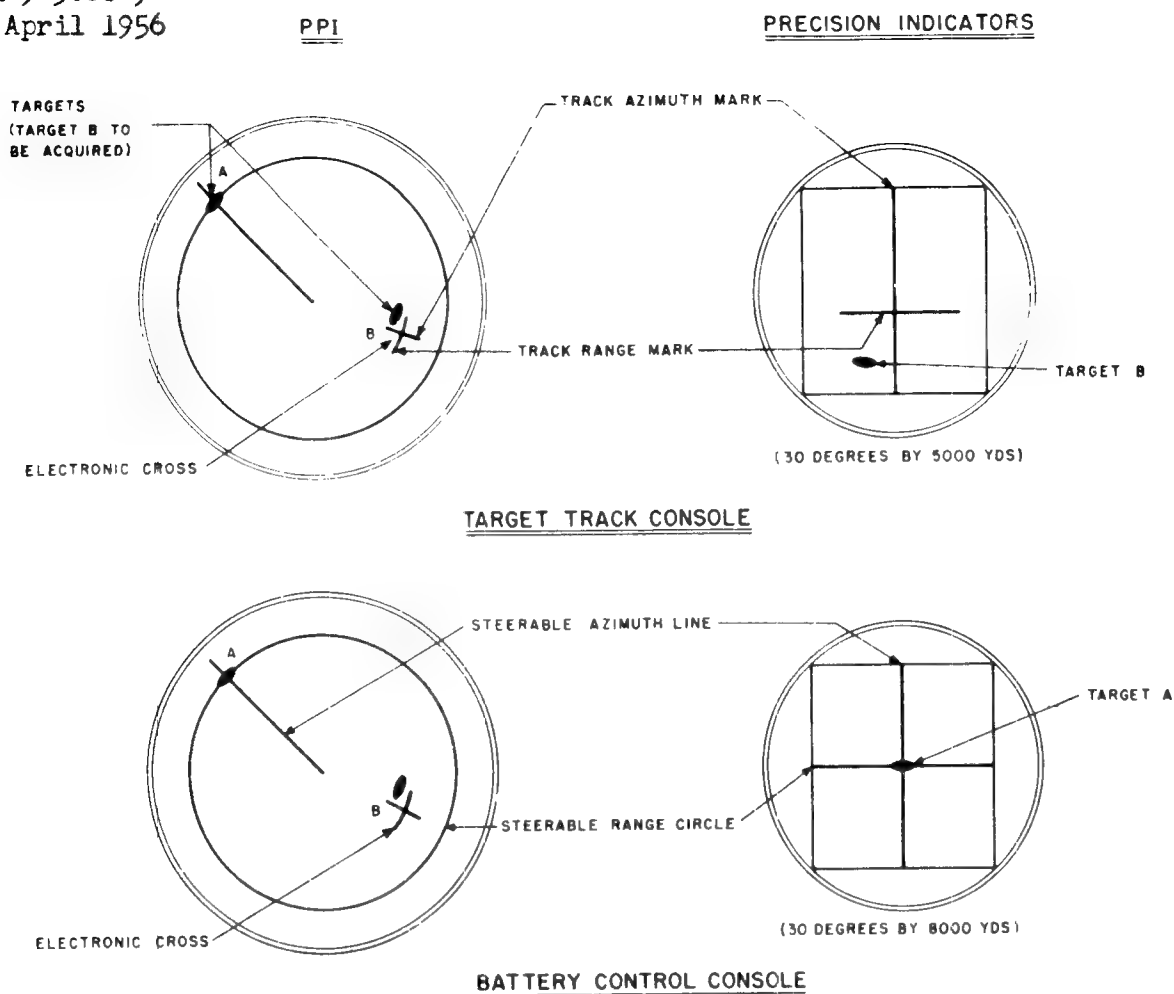


Figure 2. Acquisition system presentations.

b. Display (fig 2). Four indicator tubes are included in the acquisition presentation system. Each console (acquisition radar control console and target-tracking radar control console) contains one PPI and one B-scope precision indicator. The sweeps on the PPI tubes display acquisition radar echoes in polar coordinates at ranges of either 60,000 yards or 120,000 yards as determined by switches located at each PPI. The precision indicators show expanded acquisition echoes in rectangular coordinates; the precision indicator in the battery control console shows an expanded area 30° in azimuth by 8,000 yards in range about the intersection of the steerable azimuth line and the range circle. The precision indicator in the radar control trailer shows an expanded area of 30° in azimuth by 5,000 yards in range about the electronic cross. The vertical line in the precision indicator in the battery control trailer represents the steerable azimuth line, and the horizontal line represents the range circle; in the precision indicator in the radar control trailer, the vertical line represents the target-tracking radar azimuth setting and the horizontal line represents the target-tracking range

setting. With the precision indicators, it is possible to distinguish between two targets separated by 200 yards in range or 1° in azimuth.

c. Video. The video image presented on the acquisition indicators are acquisition video signals only; these signals may be either MTI video or bypass video. The precision indicator tubes as well as the PPI's are intensity modulated, and the video image appears as an intensified portion of the sweep.

d. Range circle. The range circle appears on both PPI's at a range determined by the RANGE handwheel on the battery control console. On the precision indicator in the battery control trailer the range circle appears as a horizontal line bisecting the face of the tube. The range circle is used to measure the approximate range to a target and to designate the range of a selected target.

e. Steerable azimuth line. This signal appears as a radial line on the face of the PPI's and as a vertical line bisecting the display on the precision indicator in the battery control trailer. It appears once per revolution of the acquisition antenna at an azimuth determined by the setting of the azimuth control knob, located in the battery control console. It is used to determine the approximate azimuth to a target and to designate the azimuth of a selected target. When the position of the azimuth line is changed, the azimuth line ring depress switch located on the battery control console is depressed. When this ring switch is depressed, the sweep on the PPI in the battery control trailer no longer rotates in synchronism with the acquisition antenna. Instead, it is deflected in a direction determined by a line slew resolver. The rotor of the resolver is connected mechanically to the azimuth line control on the battery control console. The steerable azimuth line may then be adjusted by rotating the AZIMUTH control (the range circle at this time appears as a bright spot on the steerable azimuth line). The azimuth line may be rotated without depressing the ring depress switch. However, it flashes only as the acquisition antenna sweeps by its azimuth, a much speedier location of the steerable azimuth line may be made by first pressing the ring depress switch, blanking from the scope all but the azimuth line, and then moving it accurately over the selected target.

f. Electronic cross. This element of the display is produced by four signals generated in circuits associated with the target-tracking radar. Its purpose is to indicate the position (in azimuth and range) of the target being tracked by the target-tracking radar. This is accomplished by causing the electronic cross to be generated at an azimuth and range which corresponds to the settings of the azimuth and range servo systems in the target-tracking radar. This signal is composed of a 5,000-yard radial line and a 10° (180-mil) arc.

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CHAPTER 2

SYNCHRONIZING SYSTEM

Section I. CHANNELS

14. INTRODUCTION

The synchronizing system (fig 3) is made up of three main channels whose outputs are the preknock pulse, sync pulse, and test pulse.

a. Preknock channel. The preknock channel consists of a blocking oscillator with an associated amplifier. An INTERNAL-AUTOMATIC switch determines whether the oscillator operates at its own free-running rate of 1,000 pulses per second or is triggered into operation by the AUTOSYNC pulse from the MTI system. When the switch is positioned to INTERNAL, the AUTOSYNC pulse is cut out of the circuit thus permitting the blocking oscillator to oscillate at its normal rate of approximately 930 pulses per second. When the switch is positioned to AUTOMATIC, the AUTOSYNC pulse is effective and triggers the oscillator just prior to the time it would trigger itself, producing a frequency slightly higher than the free-running frequency. The MTI system requires an extremely accurate period between pulses for proper functioning, as explained in chapter 6. The AUTOSYNC pulse, which originates in the MTI system, is timed so as to have this same critical period. Therefore, the AUTOSYNC pulse automatically locks the period of the blocking oscillator to the period required by the MTI system. Each output preknock pulse is sent to four external locations:

- (1) Target-tracking radar.
- (2) Acquisition MTI system.
- (3) Acquisition range system.
- (4) Acquisition presentation system.

The target-tracking synchronizer system receives each preknock pulse and is, therefore, synchronized with the acquisition operation. Functioning of the MTI and range systems is initiated, and sweep voltages are activated in the presentation system by the preknock pulse. The preknock is a positive 2-microsecond, 40-volt pulse.

b. Sync channel. The sync channel consists of amplifiers, a multivibrator, and a blocking oscillator. Each preknock pulse is fed down to this channel where it is delayed for 23.5 microseconds and, in effect, emerges from the blocking oscillator as the sync pulse, a positive 2-microsecond, 40-volt pulse. This time interval between the preknock and the sync pulse is necessary for proper range determination in the target-tracking ranging system. The operation of this system is closely allied with that of the acquisition radar concerning time occurrence of pulses. The output sync pulse triggers the magnetron within the transmitter system.

c. Test channel. The test channel has an amplifier and a blocking oscillator. The oscillator receives the sync pulse and emits a positive 7- to 9-microsecond, 6-volt pulse. Through the test pulse, a check is available on the operation of the MTI system.

15. OPERATION (TM 9-5000-26, page 198)

a. Over-all operation. For proper operation of the acquisition pulse synchronizer, the free-running period of blocking oscillator V2 must be slightly greater than that of the time delay of the delay cell. For proper operation of the target pulse synchronizer, the free-running period of blocking oscillator V2 must be slightly greater than the free-running period of the acquisition pulse synchronizer. The time delay varies in different sets, and the delay of each cell is stamped on the cell housing. Two adjustments in each blocking oscillator circuit may be made to fix its free-running frequency. If the free-running frequency is improperly adjusted, double triggering can occur. The adjustment procedure is outlined in the field adjustment and orientation text (TM 9-5000-23) of the Nike ground guidance system.

b. Amplifier VI. The amplifier may be considered a driver whose output is inductively coupled to the blocking oscillator. The position of the INT-AUTO switch, S1, determines whether the automatic-synchronizing pulse will be applied to the amplifier. In the AUTO position, switch S1 applies the automatic-synchronizing pulse through the amplifier and triggers the blocking oscillator. With switch S1 open, in the INT position, no signal is applied to the amplifier, and the blocking oscillator runs free. The amplifier, though energized with the proper d-c voltages, has no signal applied to its grid when S1 is in the INT position. The acquisition synchronizer will always be in AUTO during MTI operation. The target synchronizer will always be in AUTO except when operating the target radar without the acquisition radar operating during testing and alinement procedures.

c. Pulse amplifier V3A. Tube V3A is a pulse amplifier which is biased below cutoff and which will conduct only when the positive output of the blocking oscillator is applied. The output of the pulse amplifier is used to trigger the sync delay network.

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d. Sync delay network. This delay circuit consists of multivibrator V4 and delay network Z1. The multivibrator is triggered by the output of V3A, and its output cuts off diode switch tube V3B. The charge path of Z1 determines the length of time the switch tube will remain cut off. This time can be determined by the setting of the sync delay potentiometer which is normally set for 23.5 microseconds. Multivibrator V4 establishes a charge path for network Z1 for a time in excess of 23.5 microseconds.

e. Switch tube V3B. Switch tube V3B, a triode connected as a diode, normally conducts until the multivibrator output cuts it off. It remains cut off for a period of time established by the charge path of delay network Z1. The output signal, a negative 23.5-microsecond pulse beginning at preknock time, is applied directly to the grid of V5A.

f. Amplifiers V5A and V5B. Tubes V5A and V5B amplify the small output of the switch tube to an amplitude sufficient to trigger the blocking oscillator. Pulse transformer T3 in the plate circuit of V5B acts to differentiate the output of V5B.

g. General. Since the circuitry of the target and acquisition synchronizers is identical, only the acquisition synchronizer will be considered in the block diagram and circuitry discussions. One difference should be noted. The acquisition synchronizer has as an input the 10-volt, 1-microsecond, automatic synchronizing pulse. The target synchronizer utilizes the 40-volt, 2-microsecond, acquisition preknock pulse for the same purpose. In this discussion, when the automatic synchronizing pulse is mentioned, acquisition preknock must be substituted when considering the target synchronizer. Much of the discussion of this special text will also apply to the circuitry of the pulse generator in the missile radar. This unit serves the same purpose as the synchronizer of the acquisition and target radars but is different due to the variable pulse recurrence frequency of the missile radar and the necessity of having both a positive and a negative synchronizing pulse output.

16. BLOCK DIAGRAM (fig 3)

a. Blocking oscillator V2A. This stage is a single swing blocking oscillator whose synchronized frequency is approximately 1,000 pulses per second and whose free-running frequency is slightly lower. The free-running frequency is determined by the circuit constants of the blocking oscillator. During automatic synchronization the oscillator is driven by amplifier V1, which has applied to it an automatic synchronizing pulse from the MTI system. For proper operation of the blocking oscillator, the automatic synchronizing frequency must be higher than the free-running frequency. With switch S1 closed, the positive automatic synchronizing pulse is applied to the grid of amplifier tube V1.

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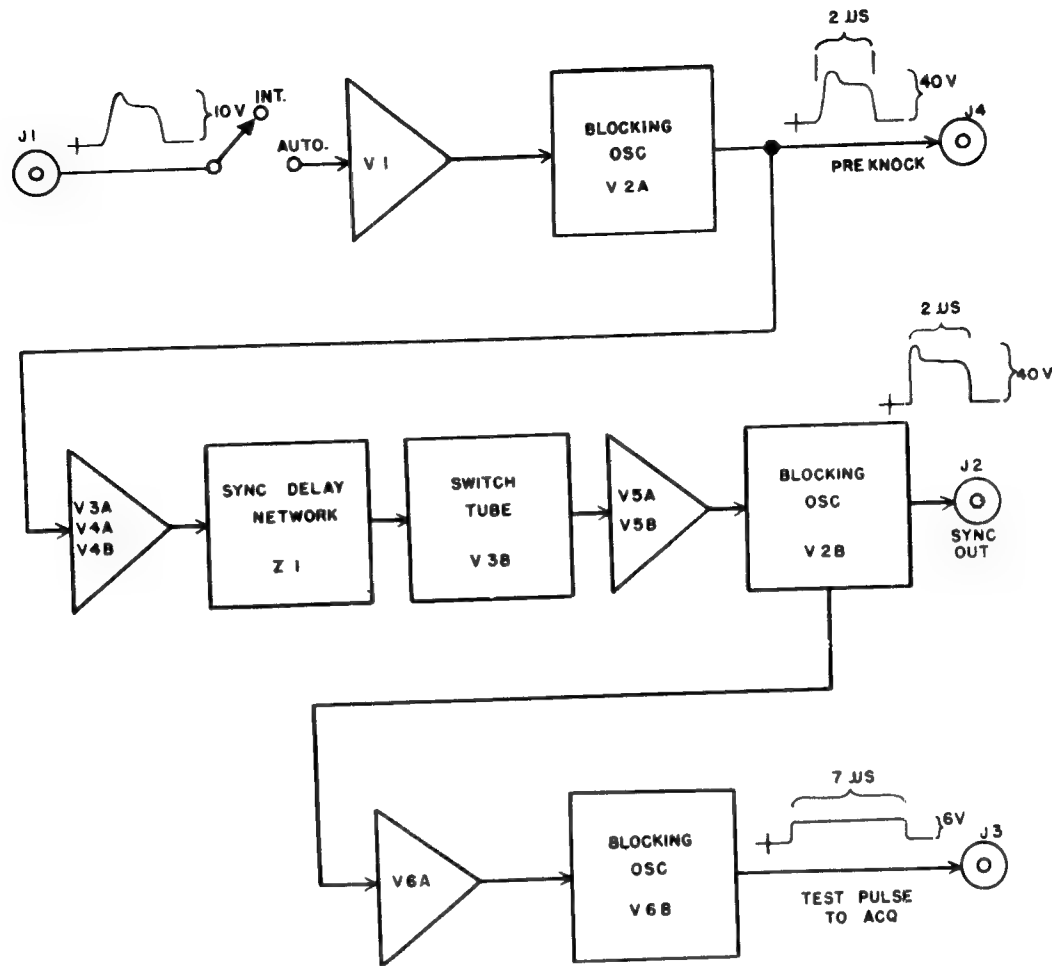


Figure 3. Acquisition synchronizer, block diagram.

b. Free-running frequency. The amplifier triggers the blocking oscillator just before the free-running repetition period is over: thus, the frequency of the blocking oscillator of the acquisition synchronizer is controlled by the automatic synchronizing pulse. The blocking oscillator output is the acquisition preknock pulse which is applied to the grid of V1 of the target synchronizer. This amplifier triggers the blocking oscillator just before its free-running repetition period is over, and the frequency of the blocking oscillator of the target synchronizer is thus controlled by the acquisition preknock pulse. The discharge path of the capacitors C4 and C15 is from the upper plates of the capacitors, through resistors R11 to R7, R44, R6, and R5, the power supply, cathode inductor T2, to the lower plates of the capacitors. Therefore, the R-C time constant of these capacitors and resistors determines the free-running frequency of the blocking oscillator. After the capacitors have discharged sufficiently to permit the tube to conduct, the regenerative cycle is repeated. The

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output signal, developed across the cathode inductor, is the result of the change in the plate current which passes through it. The signal is a 2-microsecond positive pulse with an amplitude of approximately 40 volts. The time consumed by the rise and decay of the plate current (pulse width) is determined mainly by the inductance of the transformers. The 40-volt synchronizing pulse is applied to V6A, which, acting as a driver for the blocking oscillator V6B, triggers the stage. The output of the stage is a positive, 6-volt, 7- to 9-microsecond pulse. The plate supply voltage for this stage is obtained through TEST PULSE switch S2 on the carrier generator chassis. The voltage is not applied unless the MTI is being aligned and the TEST PULSE switch is in the ON position. This stage is not used in the target synchronizer.

Section II. SYNCHRONIZING OPERATION

17. BLOCKING OSCILLATOR V2A (TM 9-5000-26, page 198)

a. After the initial application of plate voltage to the stage, plate current flows. This current flow causes a magnetic field to be developed about the plate winding of T1. The expanding field cuts the conductors in the grid winding of the transformer, inducing a voltage across it. The transformer secondary is terminated at the oscillator grid in a manner which drives the grid positive at this time. The blocking oscillator plate current increases because of the increasing potential at the grid. This regenerative action causes plate current to rise quickly to saturation. During this action grid current is drawn, and capacitors C4 and C15 become charged. The charge path of the capacitors is from the lower plates of the capacitors, from cathode to grid of V2A, through the grid winding of T1 and back to the upper plates of the capacitors.

b. When the blocking oscillator reaches saturation, plate current ceases to increase, and the magnetic field about the plate winding of T1 no longer expands. Since the voltage across the grid winding is a function of an expanding magnetic field, no voltage will be induced in the grid winding of T1. This drop of potential at the grid results in a decrease in plate current. Since the magnetic field about an inductor is a function of current through the inductor, the field about the plate winding will begin to collapse because of the decrease in current through the winding. This collapsing field induces a voltage across the grid winding of the reverse polarity, driving the grid in a negative direction. The stage is driven rapidly below cutoff, and plate current ceases. The negative potential of the charged capacitors is applied to the grid through the grid winding of T1. The tube remains cut off until capacitors C4 and C15 discharge sufficiently to reduce the bias to the point where the tube may conduct again. The discharge path of the capacitors C4 and C15 is from the upper plates of the capacitors, through resistors R11 to R7, R44, R6, and R5, the power supply, cathode inductor T2, to the lower plates of the capacitors. Therefore, the R-C time constant of these

capacitors and resistors determines the free-running frequency of the blocking oscillator. After the capacitors have discharged sufficiently to permit the tube to conduct, the regenerative cycle is repeated. The output signal, developed across the cathode inductor, is the result of the change in the plate current which passes through it. The signal is a 2-microsecond positive pulse with an amplitude of approximately 40 volts. The time consumed by the rise and decay of the plate current (pulse width) is determined mainly by the inductance of the transformers. The time constant of capacitors C4 and C15 and resistors R5 to R11 determines the length of time the blocking-oscillator tube is held below cutoff, and this determines the free-running pulse recurrence frequency.

c. The free-running frequency is somewhat lower than 1,000 pps; therefore, each period has a duration greater than 1,000 microseconds. This period may be varied within moderate limits by adjustment of R6, the fine adjustment. The coarse adjustment is fixed by the position of the jumper on the resistive network containing resistors R7 through R11. Capacitor C3B and resistor R5, a decoupling network, lower the plate voltage and hold it constant at the proper level.

d. A positive, 40-volt, 2-microsecond signal is developed at the cathode inductance tap. An inductive load is used in order to preserve the steepness of the waveform by offering a high impedance to the high frequencies contained in the pulse signal. The output is tapped at T2 to match the characteristic impedance of the output cable. The cable is terminated in its characteristic impedance. A transmission line is terminated by a resistor equal to the characteristic impedance of the line so that all energy flowing down the line is absorbed by the load and no reflections occur. Reflections would cause standing waves on the transmission line and would result in losses.

18. AMPLIFIER V1

The pentode amplifier is biased by R3 and C2 and is normally conducting. C1, C3A, R2, and R4 provide the elements of the tube with the proper d-c operating voltages and bypass any voltage variations to ground. R2, in the grid circuit, matches the impedance of the input cable. S1 permits application of the automatic-synchronizing pulse to the grid of the stage. When the INT-AUTO switch is in the INT position, no signal is applied to the stage. With the switch in the AUTO position, the positive automatic-synchronizing pulse of 1-microsecond duration and 10-volt amplitude is applied to the grid. This signal at the grid results in an increase in plate current which will induce across the grid winding of T1 a voltage which will be positive at the grid of V2A. The control grid is driven above cutoff, and the blocking oscillator conducts. Therefore, as the result of the application of the automatic-synchronizing pulse to amplifier V1, the blocking oscillator is triggered at a frequency which is slightly higher than the free-running frequency.

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19. PULSE AMPLIFIER V3A

This stage is normally biased below cutoff by a negative 27.5-volt potential obtained from voltage divider R12 and R13, which is connected between 250 volts and ground. The positive signal from the blocking oscillator, the same as the preknock pulse output, is coupled through capacitor C5 and is developed across the voltage divider, causing V3A to conduct. The negative-going plate waveform is coupled directly to the sync delay network.

20. SYNC DELAY NETWORK

a. Tube V4A is normally cut off by the bias of -21 volts established on its grid by voltage divider R16 and R17 from ground to -250 volts. Tube V4B is conducting due to +250 volts through resistor R22 to its grid. The capacitor of the Z1 network will be charged to approximately 245 volts with +250 volts on terminal 1 of Z1, since neither V3A nor V4A is conducting, and approximately 5 volts on terminal 3 of Z1 due to the conduction of V3B.

b. The conduction of V3A during the preknock pulse input triggers multi-vibrator V4. The drop at the plate of V3A and V4A is coupled across C7 to the grid of V4B, decreasing the conduction of V4B, causing a rise in potential at its plate. This rise is coupled across C6 to the grid of V4A, allowing V4A to conduct, decreasing the potential at the plate of V4A even more. This action soon causes V4B to be completely cut off and V4A to be conducting. This state will continue until V4B is again able to conduct due to its grid potential rising. This rise is the result of decreased conduction in V4A, causing a rise in plate potential. C6 charging toward -22 volts causes the decreased conduction in V4A.

c. The drop at the plates of V3A and V4A is also coupled across the capacitor of Z1, cutting off tube V3B. V3B will remain cut off for 23.5 microseconds until terminal 3 rises from some negative potential, established by the drop on the plate of V4A, to zero volts. The charge path which establishes this time at 23.5 microseconds is from the right plate of the capacitor through the resistor of network Z1, sync delay R21, the power supply and tube V4A. In order to have control of the time during which V3B is cut off by sync delay potentiometer R21, this charge path must be available. Since V4A is in this path, V4A must conduct for a period of time in excess of 23.5 microseconds. The time which V4A conducts is determined by the charge path of C6 through R18-R20, power supply, R16, R17, and cathode to grid in parallel.

d. Network Z1 is contained in a sealed oven, heated by the 6.3-volt filament voltage, to prevent variations of air temperature from causing variations of the time delay between the preknock pulse and the synchronizing pulse. The procedure for adjusting R21, SYNC DELAY, to establish the 23.5-microsecond time delay is covered in the field adjustment procedure.

21. SWITCH TUBE V3B

Switch tube V3B is a triode connected as a diode. The normally conducting tube is cut off at preknock time and remains cut off for 23.5 microseconds, as established in the discussion of the sync delay network. A negative pulse, approximately 5 volts in amplitude and 23.5 microseconds in duration, will appear across its cathode resistor. This signal is coupled directly to the grid of amplifier V5A.

22. AMPLIFIERS V5A and V5B

This circuit amplifies and differentiates the output from the switch tube to an amplitude sufficient to trigger blocking oscillator V2B. V5A amplifies the negative 23.5-microsecond pulse, and a positive pulse of greater amplitude appears at its plate. V5B may be considered as the driver for blocking oscillator V2B and is normally cut off by a negative 21-volt potential appearing at its grid established by voltage divider R27 and R28 between -250 volts and ground. The transformer plate load differentiates the output signal. The output signal monitored at pin 5 of transformer T3 is a negative pulse at preknock time followed 23.5 microseconds later by a positive pulse. This positive pulse triggers blocking oscillator V2B.

23. BLOCKING OSCILLATOR V2B

a. Operation of this circuit is very similar to that of blocking oscillator V2A. However, this circuit cannot run free. V2B is normally cut off by a bias established by a voltage divider consisting of R32, R33, and R30 between -250 volts and ground. In order to make this blocking oscillator operate, a positive trigger pulse must appear on the grid. Transformer T3 is so connected that a negative pulse at pin 5 will cause a negative pulse at pin 3. A negative pulse on the grid can have no effect since V2B is already cut off. However, the positive pulse will cause the grid to rise and begin the regenerative action characteristic of blocking oscillator operation. When V2B reaches saturation, the process reverses itself rapidly, cutting the stage off. It remains cut off until another positive pulse appears at the grid.

b. The output signal developed across the cathode resistor is the result of the change in plate current which passes through it. The signal is a 2-microsecond positive pulse with amplitude of approximately 40 volts, following preknock time by 23.5 microseconds. It is called the synchronizing pulse. This pulse is used as the output from J4 and is used to trigger the test pulse blocking oscillator in the acquisition synchronizer only.

24. AMPLIFIER AND TEST PULSE BLOCKING OSCILLATOR V6

a. V6B is the blocking oscillator and is driven by V6A. Both sections are normally held cut off by negative voltages applied to the grids. A negative 13-volt biasing voltage is developed by voltage divider network R34 and R35 and applied to the driver grid; a negative 15-volt biasing voltage is developed by R37 and R38 and applied to the blocking-oscillator grid. The synchronizing pulse is coupled through C12, developed across R34 and R35, and applied to the grid of the driver triode. This positive signal causes current to flow through the plate winding of T4. A positive voltage is induced across the grid winding and is applied to the grid of the blocking oscillator. The cyclic action of this blocking oscillator is practically identical to that of V2B. The output signal is developed across R40 and is applied to Jack J3. It is a positive, 6-volt, 7- to 9-microsecond pulse, which is sent to the carrier generator for adjustment of the MTI system. R36 is placed in the unused winding of the transformer to damp oscillations which may occur because of the inductance and distributed capacitance of the winding.

b. TEST PULSE switch S2, when in the ON position, applies plate voltage to the blocking oscillator and connects the test pulse to the appropriate circuit in the carrier generator. Switch S2 is located on the carrier generator chassis. The test pulse blocking oscillator on the target synchronizer is on the chassis only to permit interchangeability of acquisition and target pulse synchronizers.

25. OUTPUT SIGNAL DISTRIBUTION

a. The acquisition synchronizing pulse is applied to the following:

- (1) The acquisition trigger generator, where it triggers a single-swing blocking oscillator.
- (2) The IFF equipment.

b. The acquisition preknock pulse is applied to the following:

- (1) The target synchronizer, where it triggers the blocking oscillator, synchronizing zero time of the target radar with zero time of the acquisition radar.

- (2) The carrier generator, where it modulates a 15-megacycle signal and is delayed by a quartz delay cell and associated components. It is then applied as the automatic synchronizing pulse at the input to the acquisition pulse synchronizer.
- (3) The switching and mixing unit, where it triggers a one-shot multivibrator whose output signal permits gating of the MTI video for presentation on the acquisition scopes.
- (4) The acquisition and IFF control panel, where it triggers a one-shot multivibrator which produces a signal used as a variable bias for the acquisition receiver i-f preamplifier when sensitivity time control (STC) is used.
- (5) Two PPI's, where it triggers cathode-coupled, single-shot multivibrators, the outputs of which cause the PPI sweeps to be generated for ranges of 60,000 or 120,000 yards.
- (6) The two mark generators, one in each precision indicator, where the pulse activates circuits that generate acquisition and tracking azimuth marks and gates.
- (7) The acquisition range unit, where a phantastron circuit which controls the time at which the acquisition range mark and gate are generated is triggered.

c. The test pulse is applied to V3, the 15-megacycle modulator tube of the carrier generator, in place of MTI video during adjustment of the MTI system. The delayed and nondelayed video outputs of the MTI system, the gain of the two channels, must be equal so that these outputs will cancel. The test pulse is used in adjusting the gain and delay of the two channels so that total cancellation will take place.

d. The target synchronizing pulse is applied to:

- (1) The target trigger generator, where it triggers a single-swing blocking oscillator.
- (2) The i-f test unit, where it triggers circuits that develop an i-f pulse for adjusting the target receiving system.

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- e. The target preknock pulse is applied to:
- (1) The range unit assembly, where it triggers circuits that develop the tracking range mark, tracking range gate, and the acquisition tracking range mark.
 - (2) The relay amplifier, where the pulse triggers a flip-flop multivibrator and thereby makes possible a comparison of range settings of the acquisition and tracking radars. Slewing of the tracking range to the acquisition range is dependent upon this comparison.
 - (3) Three target-tracking sweep generators, where sweeps for the target-tracking indicator are initiated.
 - (4) The video error signal panel, where it triggers circuits to permit pip matching on the target-tracking indicators.
 - (5) The test delay unit, where it triggers circuits that develop an artificial target which may be varied in range to check and adjust various circuits of the target radar.

CHAPTER 3

TRANSMITTER SYSTEM

Section I. TRANSMITTER COMPONENTS

26. INTRODUCTION (TM 9-5000-26, page 195)

The acquisition transmitting system is located in the acquisition antenna assembly and consists of the trigger generator, modulator, pulse transformer, and magnetron oscillator with its tuning drive. Whenever a sync pulse is received by the transmitter system the magnetron produces a high-powered pulse of r-f energy.

a. Trigger generator. The purpose of the trigger generator unit is to amplify and expand the sync pulse from the synchronizing system. The positive 40-volt, 2-microsecond sync pulse from the sync system is expanded to a positive 850-volt, 2-microsecond trigger pulse, which is used to fire the thyatron in the modulator unit.

b. Modulator. The purpose of the modulator is to increase the positive 850-volt, 2-microsecond trigger pulse input to a 6,000-volt to 6,500-volt 1.3-microsecond pulse which is fed to the pulse transformer. The modulator consists of:

- (1) A hydrogen thyatron which amplifies the trigger pulse input.
- (2) A pulse-forming network which shapes the modulator pulse to the desired 1.3 microseconds, with a flat top and steep leading and trailing edges.
- (3) A special network which acts to shape the leading edge of the modulation pulse for proper firing of the magnetron. While the modulation pulse rises steeply, this network puts a step in the leading edge of the wave-front. This step occurs at the conduction level of the magnetron and prevents the full modulation voltage from being applied before the magnetron has a chance to start oscillating. Without a step in the leading edge of the modulation pulse, the high modulation voltage would suddenly be applied across the magnetron and result in arcing and sporadic operation.

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c. Pulse transformer. The purpose of the pulse transformer is to step up the 6,000-volt to 6,500-volt pulse to 40 to 45 kilovolts as required by the magnetron.

d. Magnetron. The magnetron tube in the acquisition r-f coupler oscillates at a frequency which may be varied from 3,100 to 3,500 megacycles (10-cm band). The minimum peak power over the tuning range is 500 kw. The magnetron is made to oscillate for about 1.3 microseconds by the 6- to 6.5-kv modulation pulse from the acquisition modulator (after the pulse transformer steps the pulse up to 40 to 45 kv). When the magnetron oscillates, the high voltage r-f energy is conducted by the acquisition waveguide assembly to the antenna, where it is beamed into space. The receiver is protected from the high-voltage pulses by a TR (transmit-receiver) tube, and the magnetron is isolated from the received echoes by two ATR (antitransmit-receive) tubes. When the magnetron has warmed up and starts to transmit, a relay is energized which reduces the heater current to prevent overheating of the magnetron filament. This relay also supplies current to the automatic frequency control motor. The frequency of the magnetron can be changed remotely from the acquisition control panel in the battery control trailer or locally from the acquisition r-f coupler in the acquisition antenna assembly. The magnetron frequency is changed by changing the size of the resonant cavities and thereby their resonant frequency. This is done by sliding tuning plugs into the cavities aligned carefully so that they enter and leave the cavities without shorting to the sides. The tuning is accomplished by a tuning motor which utilizes 120-volt, 400-cycle, a-c power and can turn a worm gear against the calibrated head of the magnetron in either direction. The head turns a screw which slides the block holding the tuning plugs in or out, depending on the direction of rotation. The tuning plugs thus enter or leave the magnetron cavities, reducing or increasing the size of the cavities and thereby raising or lowering the frequency at which they resonate. This tuning motor also turns the tap on a potentiometer which indicates the frequency on the frequency meter of the acquisition control panel.

27. TRANSMITTER OPERATIONAL DETAILS

a. Trigger generator channel. The trigger generator receives the sync pulse from the pulse synchronizer. This positive 40-volt, 2-microsecond pulse is applied to V1A, which amplifies and inverts it. The negative pulse at the output of V1A is used to trigger the single-swing blocking oscillator V1B. This oscillator is biased by -27 volts, which may be tested at TP1. The output from the blocking oscillator is a positive 230-volt, 5-microsecond pulse. This pulse is applied to the grid of thyratron V2. Before this pulse arrives, the pulse-forming network has been charged to approximately 600 volts by d-c resonant charging. The trigger pulse of V2 causes Z1 to discharge through the pulse transformer T3. T3 has a step-up turns ratio of 1:1.4. The output at TP2 is a positive, 2-microsecond pulse of approximately 850 volts amplitude. Due to an improper impedance match, some

energy is reflected back through T3. This energy would charge Z1 in the opposite direction, but that is prevented by the reverse current diode V3, which bypasses the reflected energy to ground.

b. Modulator channel. The modulator channel receives the 850-volt, 2-microsecond pulse from the trigger generator channel. This pulse is applied to the grid of the thyatron switch tube V4 causing it to ionize. However, before the pulse arrives, the pulse-forming network, Z6 has been charged to between 8-13 kv by d-c resonant charging. When V4 ionizes, Z6 discharges through the primary of the pulse transformer T1 in the acquisition r-f coupler. T1 steps up this pulse so that it becomes a negative 1.3-microsecond pulse of approximately 40-50 kv and applies it to the cathode of the magnetron. This pulse then is effectively plated to cathode voltage for the magnetron, and the magnetron oscillates during this 1.3-microsecond period. Due to slight impedance mismatches, the magnetron reflects back some energy which must be absorbed by the reverse current thyatron, V5, in the modulator. A meter circuit has been included to indicate the magnitude of this current which is an indication of modulator and magnetron efficiency.

c. Magnetron channel. The magnetron is a cavity-type oscillator which is tunable between 3,100-3,500 megacycles. Through the magnetron-tuning drive circuit, the magnetron is tuned by inserting plungers into the magnetron cavities in order to reduce the effects of interference from other radars and the effects of electronic warfare.

d. High-voltage V-9-111-6 power supply. The acquisition HV power supply delivers a d-c voltage between 6 and 6.5 kv and current between 0.45 and 0.75 ampere for use in the modulator. The d-c power is formed into pulses by the modulator to be applied to the magnetron.

28. ACQUISITION R-F SYSTEM (TM 9-5000-26, pages 196, 197)

a. Purpose. The purpose of the r-f system is to couple the r-f energy from the magnetron to the antenna, to radiate the r-f energy in a focused beam, and to cause this beam to scan the area around the radar.

b. Components. The r-f system consists of several components:

- (1) Ten-centimeter waveguide is used to connect all components within the r-f system.
- (2) A section of waveguide called the duplexer permits the use of a single antenna for both transmitting and receiving. This duplexer consists of the Y-junction, two ATR tubes, and a TR tube.

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- (3) An attenuator is attached to the waveguide near the magnetron to sample the output from the magnetron for use in the automatic frequency control of the receiver system.
- (4) A directional coupler attached to the waveguide between the duplexer and rotary joint is used when making frequency, power, and standing wave measurements.
- (5) A rotary joint allows the antenna to rotate in azimuth while maintaining the continuity of the waveguide.
- (6) The acquisition antenna radome houses a tripple reflection antenna. The antenna uses multiple reflections to focus the radiated energy into a beam which is approximately 25 miles wide in azimuth and is adjustable in elevation. The antenna and radome rotate 10, 20, or 30 revolutions per minute at the discretion of the operator.

Section II. FUNCTIONING

29. TRIGGER GENERATOR (TM 9-5000-26, page 199)

- a. General. The trigger generator, when triggered by the synchronizing pulse, produces a 2-microsecond, +850-volt pulse which is of sufficient amplitude to fire the hydrogen thyratron switch in the modulator channel.
- b. Sync pulse amplifier V1A. V1A acts as a class A amplifier to amplify the sync pulse. Resistors R4 and R5 form an effective plate load of 41,000 ohms. Capacitor C1 is a decoupling capacitor. Capacitor C3 and resistor R7 couple the input pulse to the grid of V1A. Resistor R6 terminates the connecting cable between the synchronizer and the trigger generator in its characteristic impedance. When the synchronizing pulse is applied, conduction increases, and a negative pulse is developed across the plate-load resistance. This drop is coupled through capacitor C4 to the plate of V1B. Capacitor C4 then begins to charge. Electrons leave the right plate of C4 and pass through the plate winding of T1 from terminal 4 to terminal 3. This current develops a magnetic field about winding 3-4. As the magnetic field builds up, it develops a voltage on the secondary of the transformer so that terminal 1 and the grid of V1B become more positive.
- c. Single-swing blocking oscillator V1B. Between synchronizing pulses, stage V1B is held below cutoff by a negative 27-volt potential applied to its grid. Resistors R1 and R2 form a voltage divider between ground and minus 250 volts. Capacitor C2 maintains a constant d-c level at the junction of R1 and R2, and is also a part of the blocking oscillator circuit. When the bias of the stage is overcome by the action of V1A, V1B conducts. The resulting flow of plate current

through the plate winding causes expansion of the magnetic field about the plate winding, and a greater voltage is induced in the grid winding. This causes the grid to go still more positive, and more current flows through the tube. A violent and almost instantaneous regenerative motion takes place, and the tube quickly reaches current saturation. At the instant saturation is reached, there is no longer an increase of current through the plate winding and, therefore, no increase in the magnetic lines of force. At that time no voltage is induced across the grid winding, and the grid potential drops sharply. The voltage induced across the grid winding is now of the opposite polarity and causes the grid to become more negative than minus 27 volts. The tube is driven rapidly to cutoff by this action. During the period of heavy conduction, the grid was driven positive with respect to the cathode, and the flow of grid current charged C2 to a potential more negative than minus 27 volts. After the tube is cut off, the capacitor discharges through R2, and the potential on grid 7 of V1 returns to minus 27 volts. The stage remains cut off by the minus 27-volt bias until the occurrence of the next synchronizing pulse. A positive 230-volt, 5-microsecond pulse is transformer-coupled to the grid of V2 by winding 5-6 of T1. This is the output of V1.

d. Hydrogen thyatron switch tube V2. This stage is a 3C45-type of hydrogen-gas-filled triode. Its function in the circuit is to act as an electronic switch. This stage requires a trigger of approximately 150 volts, rising at a ratio of approximately 100 volts per microsecond. The tube will operate over a wide range of anode (plate) voltages through which the grid has complete control. The grid must be driven to approximately 100 volts positive for the tube to conduct. The tube deionizes when the plate voltage is driven below the deionizing potential. The hydrogen is used to reduce deionization time to about 0.1 microsecond. The 3C45 is rated at 25 kilowatts. When the positive 230-volt, 5-microsecond pulse from the blocking oscillator is applied the hydrogen ionizes and the tube conducts. Tube V2 is in series with the primary of transformer T3, and Z1 provides a low impedance discharge path for the pulse-forming network. As the network discharges, a pulse is developed across the primary winding of transformer T3. When the tube fires and conducts it acts as a closed switch. When the tube is deionized it acts as an open switch.

e. Pulse-forming network Z1. The pulse-forming network is an open-ended section of artificial transmission line. Any wire-passing current has the property of inductance. Any two-wire transmission line will have distributed capacitance. There is a measurable delay in microseconds required for passage of energy from one end of a line to the other. It is possible to construct an artificial transmission line with inductance and capacitance which will have a predetermined time delay and a characteristic impedance to match loaded conditions.

f. Charging-inductor L1. In the interval between synchronizing pulses, the pulse-forming network is charged slowly through a charging inductance, L1, which is used in almost all line-type pulsers, since it has high efficiency and

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permits the pulse-forming network to be charged to a voltage nearly double that of the power supply. This process is called d-c resonant charging. During the discharge of the pulse-forming network, the inductor isolates the power supply from the pulse.

g. Resonant charging. In a series circuit consisting of L1, Z1, T3, and a power supply, when the power supply is turned on, the capacitors of Z1 begin to charge toward the 320 volts applied potential. After approximately 500 microseconds, the capacitors are charged to the supply potential. However, the current flowing in the inductances will continue to flow in the same direction for some time, even though no voltage is now applied. This current charges the capacitors to almost twice the applied voltage. After 1,000 microseconds, the capacitors will begin the discharge to 320 volts. However, at this time the hydrogen thyatron conducts, causing the capacitors to discharge instantaneously through the pulse transformer, T3.

h. Discharge of Z1 by V2's conduction. At the end of 1,000 microseconds, the gas thyatron switch tube, V2, is caused to conduct by a positive pulse from the blocking oscillator V1B. This causes the energy stored in Z1 to be discharged through T3. This results in a negative 600-volt, 2-microsecond pulse being developed across the primary of T3. T3 has a step-up ratio of 1:1.4, and inverts the pulse so that the output from the trigger generator is an 850-volt, 2-microsecond, positive pulse. The reason all the voltage is felt across T3 during discharge and apparently none across Z1 is that the impedance of Z1 is very small compared to the impedance of the primary of T3. This results in a mismatch in impedance, but in this circuit the important output is voltage and not power.

i. Operation of reverse current diode V3. When the pulse-forming network completes its discharge, the magnetic field in T3 collapses, causing a large voltage to be developed across its primary windings. This voltage will operate to charge Z1 in a direction opposite to the original charge, as shown in figure 4. If this is permitted to happen, Z1 will already have a negative charge on its capacitors when V2 deionizes. This will result in excessive charging current through L1 on the next charging cycle and perhaps burn out L1 or Z1. To prevent this negative charge from appearing on Z1, a diode is connected from the plate of V2 to ground. This diode will conduct only when the top of Z1 becomes negative with respect to ground, thus preventing a negative charge from being developed on Z1. The output signal from the trigger generator is then coupled to the grid of the modulator thyatron. This signal is a positive 2-microsecond, 850-volt pulse and may be observed at TP2 on the chassis if equipment is available that will handle this large amplitude pulse.

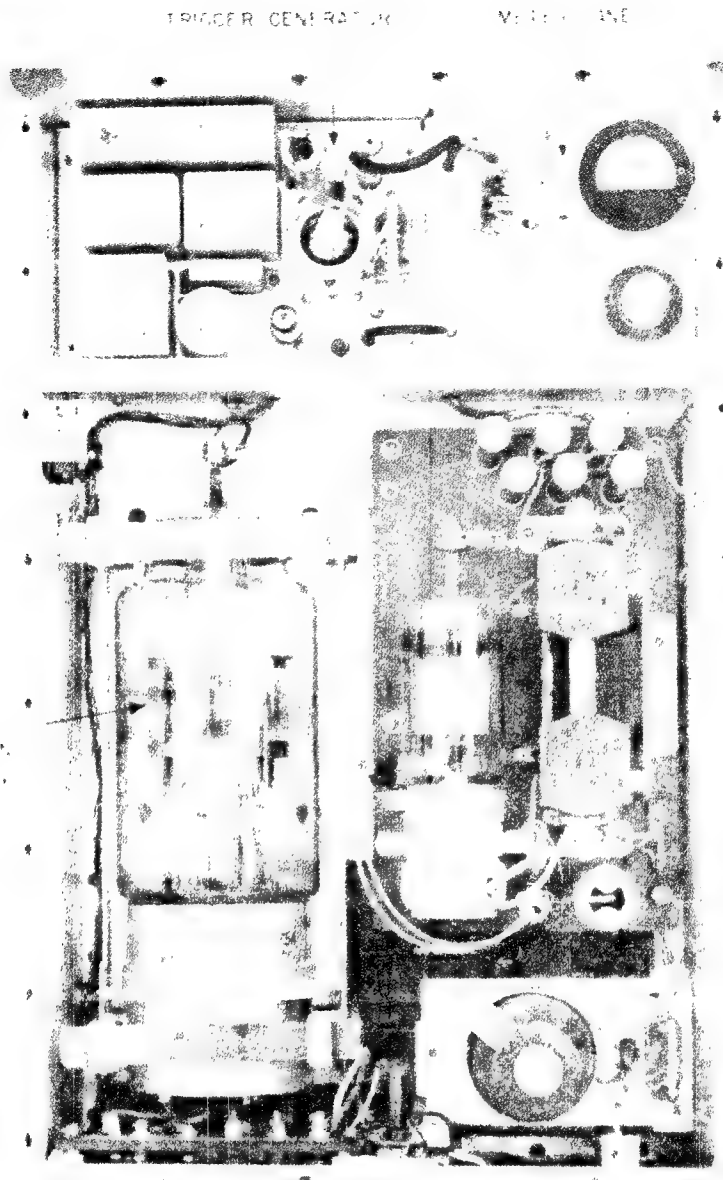


Figure 4. Modulator component arrangement.

30. MODULATOR CHANNEL (TM 9-5000-26, page 200)

a. General. The modulator is located in the acquisition modulator unit of the acquisition antenna assembly (see figure 4 for component arrangement). Narrow, flat-topped pulses that are 1.3 microseconds in duration and 6 to 6.5 kilovolts in amplitude are produced by the modulator when the output pulse of the trigger generator is applied. The modulator consists primarily of a line-type pulser, which operates in essentially the same manner as that in the trigger generator.

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- (1) The pulse-forming network Z6 is charged to from 8,000 to 13,000 volts by the high-voltage power supply and the charging inductance through the application of 4,000 to 6,500 volts for a time period of approximately 1,000 microseconds. The line is charged by d-c resonant charging. The characteristics of this pulse-forming network are such that the network discharges in 1.3 microseconds. Its characteristic impedance is 15 ohms.
- (2) Hydrogen thyatron tube V4 is the electronic switch that permits the discharge of the pulse-forming network. When the hydrogen gas is ionized, V4 conducts heavily, allowing the pulse-forming network to discharge through the primary of pulse transformer T1 and switch tube V4. Ionization is accomplished by application of the driver pulse from the trigger generator to the grid of V4.
- (3) Reverse current thyatron V5 is placed in the circuit to dissipate any reflected pulse or reverse charge across the pulse-forming network caused by an impedance mismatch of the pulse-forming network to T1. This normally results from the changing impedance of the magnetron. When a reverse voltage appears across Z6, the reverse current thyatron will fire and rapidly dissipate the charge on the pulse-forming network.
- (4) Meter M1, together with its associated switch, permits the filament voltage of the hydrogen thyatron gas capsule and the current in milliamperes through the reverse current thyatron to be measured. Relay K1, in the circuit between reverse current thyatron V5 and ground, is a protective device that removes the high voltage from the modulator in the event of excessive reverse current.

b. Resonant charging. Resonant charging of the modulator pulse-forming network follows the same theory as that covered in the trigger generator, although higher power is involved. Z6, the pulse-forming network, charges through the charging inductance L2. Normal operation would require approximately 6 kv input from the power supply. Under these conditions, Z6 will charge to 12 kv, or approximately twice the applied voltage due to resonant charging.

c. Discharge of Z6. At the end of 1,000 microseconds, the trigger pulse is applied, and the thyatron switch tube ionizes, causing Z6 to discharge through the primary of the pulse transformer T1. This time, however, proper impedance match is maintained between the pulse-forming network and the pulse transformer so that only half the voltage is felt across each of the two elements during discharge. The result is a negative, 6 kv, 1.3-microsecond pulse across the primary of the pulse transformer T1. In the modulator, impedance matching is important because of the high power involved. Small impedance mismatches result in large

amounts of reflected energy and cause high reverse current. The pulse-forming network has an impedance of 15 ohms, and the magnetron has an impedance of 800 ohms. To match these two units, the pulse transformer T1 has an impedance ratio of 15 to 800 ohms and a turns ratio of 1 to 7.3. Even with these design features, some mismatch exists. The match cannot be perfectly maintained because of the changes in magnetron impedance with magnetron age and frequency changes.

d. Reverse current circuit. The reverse current meter M1 indicates the relative mismatch existing in the system. In the case of an exceedingly good impedance match, meter M1 will read about 10 ma reverse current. Values up to 90 ma are acceptable. Values in excess of 200 ma will cause relay K1 to turn off the a-c input to the acquisition high-voltage power supply. The reverse current thyatron V5 operates to prevent the top of Z6 from going negative. This negative effect is caused by the collapse of the magnetic field in the primary of the pulse transformer T1 at the end of the discharge cycle as shown in the discussion of the trigger generator. The reverse current thyatron V5 is a self-triggering switch tube. C3 and C4 charge to the voltage on the cathode of V5 through R11. As long as the grid and cathode of V5 are at the same potential, V5 does not conduct. However, if instantaneously the cathode was made negative due to a negative charge on Z6, a voltage would develop across PC3. This voltage would cause V5 to ionize. V5 would then conduct from cathode to ground through R12, through R17, R21, R20, R3, and the parallel path formed by meter M1 and relay K1. The circuitry associated with the meter is used to smooth out the pulses of reverse current so that they can be used to operate the relay and meter. Thus C1, C6, R22, and C10 comprise a filter for the meter and relay. R3 is a 1 percent meter shunt. Relay K1 energizes and turns off the acquisition high voltage when the average reverse current exceeds 200 milliamperes.

e. Thyatron V4. When the switch associated with meter M1 is positioned so that the meter is across its terminals 3 and 9, meter M1 will read the a-c voltage applied to the gas capsule in the hydrogen thyatron V4. This voltage can be varied by adjustment of variac T2. The voltage applied to this gas capsule, the characteristic of each particular thyatron, and the temperature in the modulator will together determine the pressure of the hydrogen gas inside the thyatron tube envelope. The pressure of this gas will, in turn, determine the positive potential of the pulse that is necessary to ionize the hydrogen gas. The positive potential necessary to ionize the gas varies inversely as the pressure of the gas. The pressure of the gas varies directly as the voltage applied to the capsule. Thus, as the voltage is raised from 4 to 5 volts, gas pressure increases, and the gas will ionize with a smaller positive potential applied to the thyatron grid. Variac T2 is normally adjusted so that meter M1 reads the voltage which is stamped on the hydrogen thyatron in use. However, as the thyatron ages or as the temperature surrounding the thyatron goes to an extreme, the voltage stamped

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on the thyatron may not be the capsule voltage at which the thyatron will operate properly. The capsule voltage should be so set that the thyatron will ionize each time a pulse is delivered to it from the trigger generator and so that the thyatron will quickly cease to conduct after the trigger generator pulse passes. An indication of whether the thyatron is working properly can be gained from observing the acquisition high voltage and current meter. If the meter shows a high current and low voltage reading, the capsule voltage may be too high and should be reduced. Conversely, if the meter shows a low current coupled with a high voltage reading, the capsule voltage may be too low and should be raised. If, with the capsule voltage properly adjusted to the voltage as stamped on the thyatron, the voltage and current are not in proper relationship as read on the acquisition high voltage and current meter, the capsule voltage should be raised or lowered until they are in proper balance.

NOTE: After variac T2 has been readjusted, allow at least 10 minutes for the gas pressure in the thyatron to change to its new value before analyzing the new voltage and current readings on the acquisition high voltage and current meter.

f. Installation of the 5948/1754 thyatron. Several improvements have been made on the 5948/1754 thyatron since it was originally developed. The tube has appeared in three different physical configurations: tubes with an external hydrogen reservoir, tubes without a base, and tubes with a base. Tubes having the external hydrogen reservoir and tubes without a base probably will not be found in the equipment or as spares. They are mentioned here to illustrate their effect on the present design of equipment. The external hydrogen reservoir is contained in a separate bulb attached to the stem of the main envelope of the tube. This tube requires extreme care in handling to prevent separation of the small bulb from the stem. The tube without a base is electrically identical to the tube with the external reservoir except for connections. Each of these tubes requires a bias, which is furnished by R19, C7, C8, and C9 of the acquisition modulator. Most tubes in use today will have a metal shell base with a mounting flange. The tubes with a metal base do not require the cathode bias. The bias network is grounded by connecting a strap between terminal 6 of E1 and ground. Terminal 6 is ungrounded for all tubes without bases. Systems now leaving the factory contain JAN 5948/1754 tubes with leads having the following color code:

GRID-	Green.	HEATER-	Yellow w/black sleeve (internally connected to reservoir).
HEATER-	Yellow.	RESERVOIR-	Red w/yellow sleeve (internally connected to heater).
CATHODE-	Black.	RESERVOIR-	Red.

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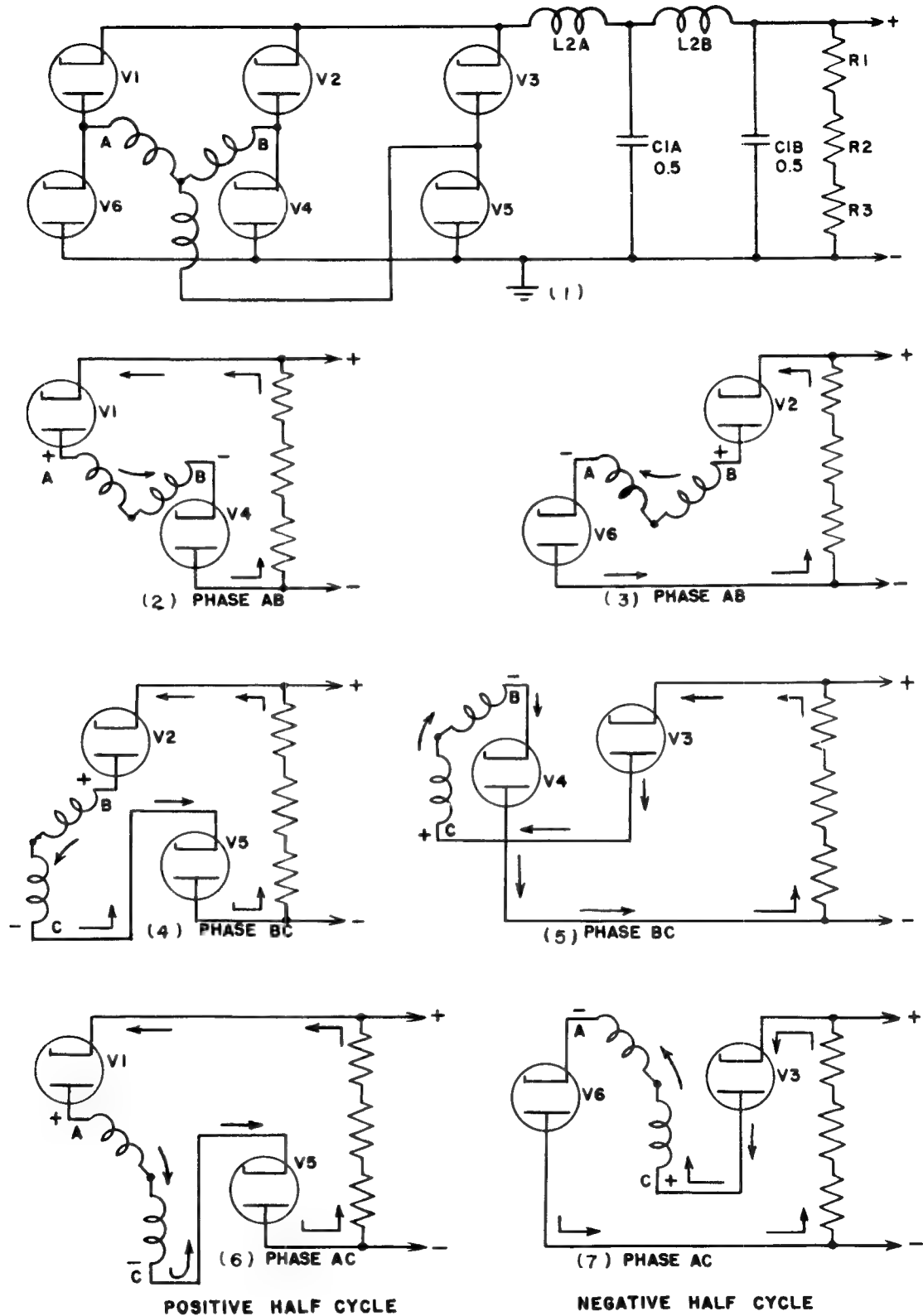


Figure 5. High-voltage power supply, operation.

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Present specifications permit reservoir voltages between 2.5-volt and 5.5-volt alternating current. Therefore, some tubes may have a THYRC VOLTS rating beyond the 5-volt full-scale range of the THYRC VOLTS meter. A future modification will be made to permit the meter to read on scale for the rated voltage of all tubes.

g. Stop network. In the plate circuit of thyatron V4 there is a damped resonant circuit consisting of L3, C2, and R18. This network is caused to oscillate by the large current pulse through it when Z6 discharges. It oscillates for 1/4 cycle and has the effect of putting a step on the leading edge of the 1.3-microsecond pulse out of the modulator. This step causes the magnetron to begin to oscillate before the high voltage part of the pulse arrives in order to prevent the magnetron from arcing by having it already oscillating when the high voltage is applied.

h. Grid and cathode circuits. Thyatron V4 receives its trigger pulse from the trigger generator through inductance L1. This inductance serves to isolate the trigger generator during the time the thyatron is ionized. Deionization of V4 is accomplished by the use of cathode resistor R19 and capacitors C7, C8, and C9. During ionization some voltage is developed across R19, which at the end of the discharge cycle acts as cathode bias and helps de-ionize V4.

Section III. ACQUISITION HIGH-VOLTAGE POWER SUPPLY

31. INTRODUCTION (TM 9-5000-26, page 201)

The purpose of the acquisition high-voltage power supply is to deliver 6 to 6.5 kilovolts at 0.45 to 0.75 ampere to the modulator of the acquisition transmitter. This voltage is adjustable to allow the magnetron to operate at reduced power when desired. The acquisition high-voltage power supply is located in the acquisition radar cabinet assembly (fig 5). Controls for this supply are located on the acquisition power control panel. Three-phase, 208-volt, 400-cycle input to the acquisition high-voltage power supply is applied through saturable reactor L1. The impedance of the saturable reactor to alternate current is determined by means of a direct current. The d-c control current for the saturable reactor is obtained from selenium rectifier CR1. The a-c voltage applied to this rectifier is obtained from the VOLTAGE ADJ variac, which is connected between phase C and neutral. A change in the adjustment of the variac will vary the input to CR1 causing a change in the value of the d-c control current applied to the saturable reactor thereby varying the impedance of the a-c windings. The a-c winding of the saturable reactor is in series with the three-phase input and the primary of the rectifier plate transformer T1. The secondary winding of the transformer T1 delivers the three-phase a-c power to a 6-tube, bridge-type rectifier. This rectifier develops d-c voltage of 6 to 6.5 kv. This voltage furnishes the high potential required for proper operation of the modulator channel.

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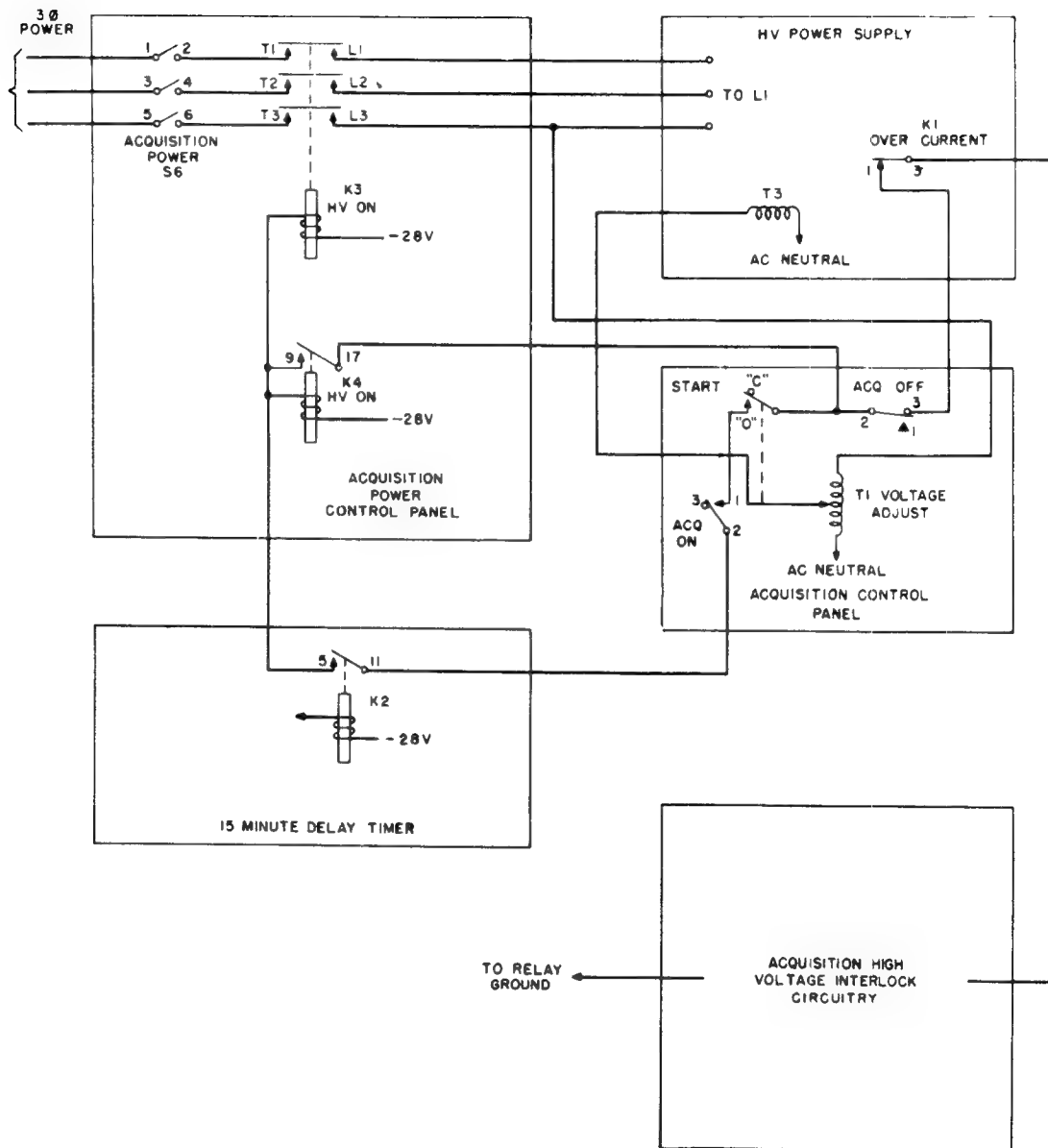


Figure 6. External control circuitry, simplified schematic.

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32. DETAILED DISCUSSION (TM 9-5000-26, page 201)

a. Control circuitry. The circuitry for applying three-phase power to the primary of transformer T1 is shown in figure 6. Three-phase power is made available in the acquisition power control panel when ACQUISITION POWER switch S6 is turned on. In order for power to be supplied through L1 to T1 in the HV power supply, HV ON relay K3 in the acquisition power control panel must be energized. The initial relay ground is applied from contacts of relay K2 in the 15-minute delay timer. However, the acquisition magnetron filaments have to be heated for 15 minutes before energizing relay K2. The contacts of K2 obtain relay ground from circuitry in the acquisition power control panel. The variable transformer on this control panel (VOLTAGE ADJUST transformer T1) must be turned fully counterclockwise to the START position so that the START sensitive switch will be closed. When the ACQ ON switch is pressed, this ground circuit is completed in this panel through the normally closed ACQ OFF switch. If all interlocks and switches are in the correct operating positions in the HV interlock circuitry, relay ground will be applied through contacts of OVERCURRENT RELAY K1, in the power supply and the above discussed circuitry to HV ON relays K3 and K4 in the acquisition power control panel. At this time, relay K4 closes holding-contacts 9 and 17 which keeps the two relays energized. Now, three-phase power is applied to L1 in the power supply, and one phase is applied to the variable transformer control in the acquisition control panel. By increasing the VOLTAGE ADJUST arm of T1 for higher voltage, more power is applied to the primary of T3 in the power supply; more d-c control current flows into saturable reactor L1, and the d-c output of the power supply increases.

If it is desired to stop acquisition magnetron operation, the ACQ OFF pushbutton is pushed on the acquisition control panel (fig 6). This action removes grounds from relays K3 and K4 in the acquisition power control panel and removes the three-phase power from the high-voltage power supply. If an excessive amount of current is drawn from the output of the high-voltage power supply, OVERCURRENT relay K1 will energize and remove ground from relays, K3 and K4 in the acquisition power control panel, which removes input power from the power supply.

b. Monitoring. Provision for monitoring is made on the acquisition high-voltage power supply. Meter M1 and MAG MA switch S1 in the control panel comprise the monitoring circuit. The switch is a center-positioning switch which, in this position, applies a portion of magnetron current to the meter. When switch S1 is pushed to one side, a voltage proportional to the high-voltage power supply output is applied to the meter from the junction of resistors R2 and R3 in the power supply. When switch S1 in the control panel is pushed to the opposite side, it applies a portion of high-voltage power supply current to meter M1 from the junction of the connection of resistor R4 and relay K1 in the power supply.

CAUTION: When a selenium rectifier is subjected to an overload, it may break down and give off extremely poisonous fumes. An indication that the rectifier has broken down is that the voltage adjust variac (MIN-MAX control) has no control over the acquisition high voltage. If this happens, the repairman should clear personnel from the van, depress the high-voltage OFF button, and allow the ventilating system to remove the poisonous fumes. In removing the faulty rectifier, the repairman should use extreme care to keep from getting the selenium residue on his skin.

c. Saturable reactor L1. The saturable reactor is a device wherein the inductance can be controlled by a direct current. L1 consists of three a-c windings and one d-c winding on the same core. The d-c winding is the control winding. The direct current through the control winding is controlled by the MIN-MAX control knob. The direct current controls the magnetization of the iron core of the reactor. The magnetization increases with an increase of current and decreases with a decrease of current. The alternating current adds to this magnetization. However, as the magnetization of the core is increased, an additional current will not produce a proportional increase in strength of the magnetic field. This is because, as the magnetizing current increases, the core becomes saturated with lines of force and its permeability decreases. Thus, at complete saturation, the inductor acts as an air core inductor, and further increase in magnetization current produces the same increase in lines of force as would be caused if the inductor had an air core. Thus, the inductance of the reactor to the alternating current varies inversely as the amplitude of the direct current control. The inductor acts as part of a voltage divider network for the a-c voltage. The path of the alternating current is from the generator through the inductor and through the primary of the high-voltage plate supply transformer. The voltage across the primary of the plate transformer can thus be controlled by controlling the impedance (inductance) of and, therefore, the voltage drop across the saturable reactor, L1.

d. Three-phase rectifier. The major component of the acquisition high-voltage power supply is the three-phase rectifier which employs six tubes in a full-wave bridge circuit. The operation of the rectifier is explained with the aid of figure 5, which shows operational diagrams of the circuit with relays and protective devices omitted. For simplicity, only the secondary windings of the three-phase, delta-to-wye, high-voltage transformer are shown. Figure 5 also contains a set of six diagrams, two for each of three phases, to show the direction of electron flow during the alternations of the a-c voltage applied to the rectifier tubes. During that part of the cycle when the voltage induced in the secondary windings of transformer T1 is such that phase A is positive with respect to phase B, the plate of V1 is made positive with respect to its cathode, and the cathode of V4 is made negative with respect to its plate. As a result, both diodes conduct, and current flows through the bleeder resistors and the load in such a direction as to produce the polarity indicated in figure 5(2). During that part of

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the cycle shown in (3) of figure 5, when phase A is negative with respect to phase B, the plate of V2 is made positive with respect to its cathode, and the cathode of V6 is made negative with respect to its plate. Both of the diodes will conduct, and current will flow through the load in the same direction as previously. The remaining diagrams of figure 5 show that the current flowing through the load as a result of phase voltages BC and AC is in the same direction. Because the voltage developed across bleeder resistors R1, R2, and R3 is contributed by all three phases, as described above, the over-all effect is the production across these resistors of a voltage with a ripple frequency of 2,400 cps. This comparatively high ripple frequency gives a high average d-c level, and less filtering is required for its removal. Each tube in the power supply handles one-third of the total current. Filtering is accomplished by L2A, L2B, C1A, and C1B. If one of the six tubes should become inoperative, the output voltage will be reduced, but the system will continue to function. Filament power is supplied by transformer T2. This transformer has four secondary windings. The filaments of V1, V2, and V3 are all at the same d-c level, but the filaments of V4, V5, and V6 are at different potentials and require individual filament supply. The four secondary windings of T2 must be well insulated from each other because of the high-potential differences present.

e. Protection devices. Overcurrent relay K1 becomes energized in the event of an overload resulting in excessive current flow in the power supply circuit. When energized, K1 opens the circuit of acquisition high-voltage relays K1 and K2, causing the removal of a-c power to the high-voltage power supply. Spark gaps TY1 and TY2 protect the current meter and the voltmeter, respectively, from damage resulting from any sudden surge of current. The door-operated switch, S1, closes when the door to the high-voltage compartment is opened. This action grounds the output of the high-voltage power supply and discharges capacitors C1A and C1B. Opening the high-voltage compartment door also opens interlock S2. The opening of this interlock deenergizes relays K1 and K2. Deenergizing K1 removes the a-c input to the power supply.

f. Metering circuits. The magnetron plate current is read on meter M1 when meter switch S8 is in position 1. The meter is shunted by a 4.99-ohm resistor R13. It is in series with an 898-ohm, current-limiting resistor R5. In addition, capacitor C3 is placed across meter M1 to short out any current surges. Filter network Z1 prevents high-frequency energy from leaking out of the high-voltage compartment in which the pulse transformer and the high-voltage leads are placed. The sparks gap at the cathode of the magnetron and the four VR tubes V6-9 at the lower end of the filter network Z1 provide low resistance paths to ground, should some part of the magnetron current path open. When an opening occurs, a large charge is built up by the high-voltage, modulation pulses across the secondary windings of the pulse transformer. If the filament leads of the magnetron should open, the spark gap arcs, preventing a static charge from

building up on the upper end of the secondary windings. If an opening occurs somewhere in the meter circuit, the four VR tubes V6-9 fire. This provides a low resistance magnetron current path so that a static charge cannot build up on the cathode of the magnetron. In this case, resistors R6-9 in the plate leads of V6-9 limit the magnetron plate current since the VR tubes maintain a 150-volt drop when conducting. Capacitors C8 and C9 are connected in parallel across the magnetron filament winding of transformer T2, while capacitors C2 and C7 are parallel and connected in series with the gate pulse line which is taken from one side of the filament winding of transformer T2. Part of the magnetron plate current comes up through resistor R36 which is connected to the gate pulse line. The voltage developed across R36 by the magnetron plate current during 1.3 microseconds of oscillation provides a gate pulse for the automatic frequency control (Acq) unit. Relay K4, energized by the magnetron plate current, applies power to the automatic frequency control tachometer motor B3.

33. MAGNETRON CHANNEL (TM 9-5000-26, pages 196, 197)

a. General. The purpose of the magnetron channel is to receive the negative 6- to 6.5-kilovolt, 1.3-microsecond pulse from the modulator and convert this into r-f energy of 10-centimeter wavelength. Associated with the magnetron are several specialized circuits. Due to the very high cost of magnetrons and their relatively short life, care is taken to protect them by various metering devices in the cathode, filament, and tuning circuits.

b. Pulse transformer T1. The pulse transformer, located in the r-f coupler, receives 6- to 6.5-kilovolt, 1.3-microsecond pulses from the modulator section. This transformer is especially designed to pass a pulse without introducing substantial change in waveform. It has a welded steel case containing oil in which the transformer windings are immersed. The transformer matches the magnetron load to the impedance of the input pulse cable from the modulator. The voltage change accomplished by T1 is a voltage step-up without inversion. This step-up makes possible the use of the comparatively low 6- to 6.5-kilovolt potential in the cable between the modulator section and the pulse transformer. This reduces the problems of high-voltage flashovers and insulation breakdown. Pulse transformer T1 has three windings. The primary winding is connected to the pulse cable. The two secondary windings, identical to each other, are connected one in each line between the filament transformer and the magnetron filaments. This is called a bifilar winding, and it is made by winding two insulated conductors side by side so that exactly the same voltage is induced in each. Since the outer case of the magnetron functions as the plate, the high voltage is more easily and safely applied to the cathode circuit. The magnetron case is grounded, and stray capacitance between plate and ground eliminated. However, the filament transformer, T2, presents a problem. If a single secondary were used on the pulse transformer, the filament transformer would also introduce additional capacitance in the pulse circuit which would seriously affect the pulse shape. A double secondary winding

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is used, therefore, and the same voltage is induced across each winding. The filament transformer is connected across the low-voltage ends of the two windings, and the magnetron filament is connected across the high-voltage ends. The voltage between the filament transformer and ground is low, and there is no need for special insulation. The capacitance introduced is negligible. The low-voltage and high-voltage ends of the pulse transformer secondaries are shorted for video frequencies by capacitors C6, C8, and C9, thus bypassing the filament secondary. Stated in another way, these capacitors allow the pulse current to divide between the two parts of the bifilar winding without affecting the heater current.

c. Magnetron cathode circuits. The cathode circuits for the magnetron are rather elaborate because the current of the magnetron is used in a variety of ways. When the magnetron fires, conduction takes place from cathode to the grounded plate. The main pulse current goes from ground through R36, developing a positive AFC gate pulse. From R36 it passes through C2 and C7 to the pulse transformer secondary, and from the bottom side of the pulse transformer back to the magnetron cathode. Because of the voltage developed across R36, a similar voltage is developed across Z1. A1 acts to filter the pulses of magnetron current so that the average value of current may be measured. When the magnetron fires, the capacitors of Z1 charge to about 30 volts. During the period between pulses, the capacitors discharge, thus providing the current for the meter circuits and relay K4. Regulator tubes V6 to V9 are inoperative except if an open occurs in the meter circuits. If such a failure occurs, these tubes will fire to prevent Z1 from burning out due to excessive voltage. The transmitter would continue to operate, but relay K4 and the two meters, M1 and M3, would not indicate magnetron current. The complete discharge path for Z1 is from Z1 to ground, from ground through R13, the meter multiplier for M1, through K4 and R50 in parallel, and back to the other side of Z1. This discharge results in a reading on M1 in the r-f coupler if the meter switch is in position 1. When this current reaches 30 ma, relay K4 closes. This same average magnetron current may be read on the radar power panel by the acquisition meter M3 with switch S5 in the center position. Relay K4 does two things. It connects motor X to the receiver tuner motor so that the AFC can tune the receiver, and it lowers the filament voltage on the magnetron. During warmup, the magnetron heater is run quite hot with 125 volts across its filament. However, when in operation, the cathode tends to overheat due to high cathode current so only a small filament current is needed. When K4 energizes, the magnetron filament ammeter M3 is connected into the circuit along with the variac T4. Both of these elements are on an upper extension of the r-f coupler motor panel. With K4 energized, the variac T4 is adjusted until meter M3 reads the specified heater current marked on the side of the magnetron. A typical value of current is 2.8 amperes.

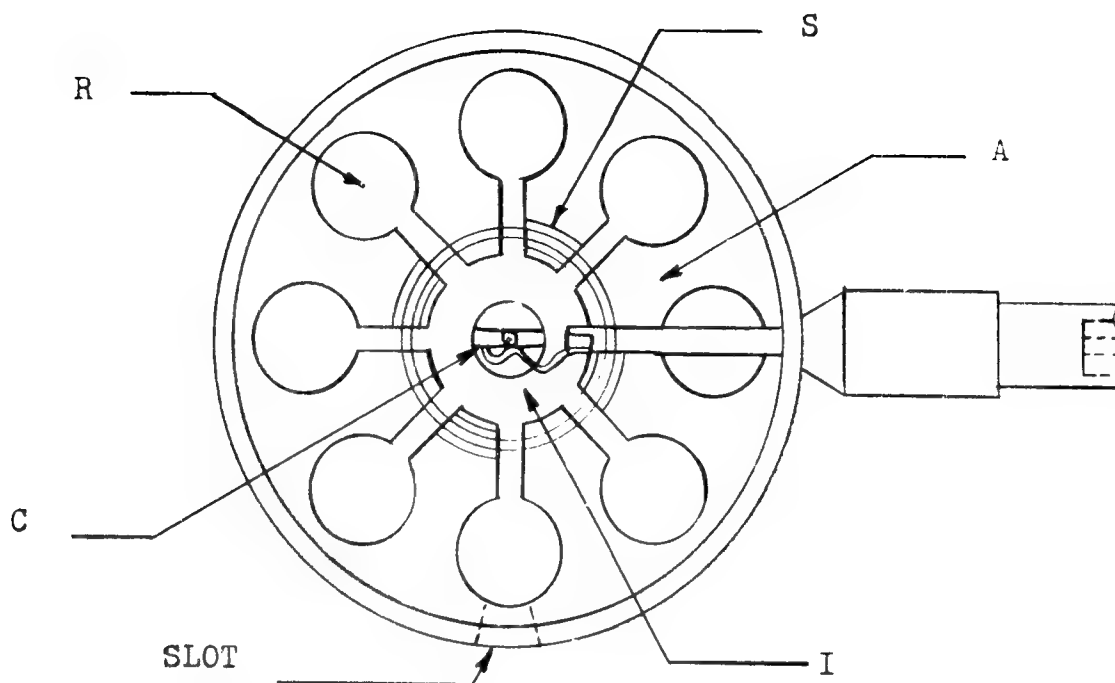


Figure 7. A typical cavity magnetron, cross-section.

d. Description of a multicavity magnetron. Magnetrons are basically self-excited oscillators whose purpose is to convert d-c input power into r-f output power. Figure 7 shows a diagrammatic sectional view of a typical magnetron. Between the cylindrical cathode C and anode block A is an interaction space I in which the conversion of d-c power takes place. A constant and nearly uniform magnetic field is maintained in this interaction space in a direction parallel to the axis of the cathode. In operation, the cathode is at a negative potential while the anode block is at ground potential. The anode block is pierced in a direction parallel to the axis of the cathode by a number of resonant cavities, R, which open into the interaction space. The inner surface of the anode, therefore, consists of alternate segments and gaps. The ends of the resonant cavities open into chambers called end spaces, through which pass lines of flux extending from one resonator to the next. The coupling between the resonators is increased by conducting bars or straps, S, which connect alternate anode segments. Energy is extracted from one resonator cavity through a small gap or slot which extends to the anode outer surface. The combination of resonant cavities, spaces, straps, and output circuit is called the resonant system. The cathode is oxide-coated and is heated indirectly by an internal heating coil of tungsten. The cathode structure is supported mechanically by an insulating material which provides anode-to-cathode isolation. The dimensions of the interaction space are designed for the

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wavelength and voltage at which the magnetron is to operate. The number of resonant cavities also influences the frequency and power of the magnetron. The resonant system shown consists of eight cavities. Each of these cavities is similar to a simple oscillating circuit consisting of lumped inductance and capacitance. The inductance of each cavity exists mainly in the circular hole, and the capacitance mainly between the parallel plane surfaces of the slot. Since the equivalent circuit of the magnetron may be represented by parallel resonant circuits equal in number to the number of cavities, it is apparent that the frequency of the magnetron will be nearly that of a single cavity. The magnetic field parallel to the axis of the cathode is produced by a permanent magnet with poles external to the magnetron tube.

e. Operation. In the operation of the magnetron, a high d-c potential is applied between cathode and anode, setting up a radial electric field. The axial magnetic field is provided by a permanent magnet. Electrons emitted from the cathode experience a force directed radially outward due to the d-c electric field, and a force perpendicular to their instantaneous direction of motion due to the magnetic field. The result is that each electron takes a path that is a series of spirals between the cathode and anode. These circling electrons excite the resonant cavities as they go by the cavity opening, resulting in oscillations in the cavity.

f. The acquisition radar magnetron. The transmitting oscillator used in the acquisition radar of the Nike I is an S-band, 5795 magnetron. The transmitted frequency is variable from 3,100 to 3,500 megacycles. This magnetron is a multicavity strapped magnetron. The theory and operating characteristics of magnetrons discussed in earlier paragraphs are directly applicable to the 5795 magnetron. Upon application of the 1.3-microsecond pulse, the magnetron oscillates at the frequency to which it is tuned. The high-voltage energy is coupled into the 10-centimeter waveguide. Power is extracted from a single resonant cavity by means of a slot which expands, with necessary discontinuities, until the dimensions correspond to those of the waveguide. Tuning plugs are inserted into each resonant cavity, and extreme care must be taken in the design and handling of the magnetron so that these plugs are not permitted to touch the sides of the cavities. Contact will cause the r-f energy to be shorted to ground. These plugs are connected to a tuning head which may be displaced by gearing to move the plugs in or out of the cavities. When the plugs are inserted further into the cavities, the inductance of each cavity is decreased and the frequency increases. The tuning head is connected by gearing and a flexible cable to the magnetron tuning drive motor so that the frequency may be changed by operating the MAGNETRON FREQUENCY switch, S4.

g. Magnetron aging. Magnetrons when received do not contain a perfect vacuum. Because of this, proper aging of the tube is necessary if satisfactory life is to be obtained. To age a new magnetron, operate it with very reduced input voltage at first and gradually build up to normal input voltage. To properly age a magnetron, raise the modulator high voltage to the point where the voltmeter fluctuates, indicating magnetron arcing. After operating at this voltage for a time, raise the voltage until the fluctuations occur again. This process is repeated until full input voltage is applied without arcing as indicated by modulator high-voltage fluctuations.

h. Operation with the 5795 full-power magnetron. An original design objective for the 5795 magnetron was to have a tube that would operate at 1-megawatt, output power level. difficulty in aging of tubes to operate at this power level was experienced by the design people at first; consequently, tubes that operating personnel may be familiar with are the 1/2-power tubes that have been manufactured during the interim. Now there are full-power (1-megawatt) magnetrons coming into service. The full-power 5795 magnetron can be identified by the words FULL POWER stamped adjacent to the serial number on the tube. This tube will operate over the assigned range of frequencies at a peak power of 1 megawatt on radars having the required modifications. All Nike systems operating satisfactorily at 1/2 power may possibly be operated at full power without difficulty, however a higher degree of performance is required from all elements in the pulsing and r-f circuits at full power than at 1/2 power. At higher power levels, the reservoir heating voltage range of the 5948/1754 thyratron is less, the alinement of the waveguide section becomes critical, and arcing may occur especially at particular frequencies on the high end of the band. Some 5948/1754 thyratrons, which perform adequately at half power may not operate at full power. In this case, refer to ordnance directives pertinent to field changes affecting full-power operation. Full-power magnetrons should be operated at a plate current of 60 ma and a heater current of 2.7 amperes to obtain a peak power output of 1 megawatt. These conditions will produce an average power output of 1,200 to 1,600 watts. At this power level and especially at high altitudes it may be necessary to use the pressurization unit to prevent arcing. If the pressurization unit is required and should fail during operation, the magnetron plate current should be reduced and the frequency changed (if permissible) to eliminate arcing. The full-power magnetron can be operated at 1/2 power by using a plate current of 33 ma and a heater current of 2.9 amperes. At this power level the use of the pressurization unit should not be required. A program for periodically conditioning spare magnetrons on a routine basis should be followed in using organizations to which the maintenance man is assigned. The service life of both 5780 tracking radar and 5795 magnetrons will be extended if they are conditioned after a period of idleness. The initial conditioning process should be as follows. The pulsed voltage should be applied gradually, allowing the tube to arc somewhat at each higher level until the arcing essentially ceases.

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Fluctuations of the magnetron current are usually an indication of arcing. This is a process of cleaning up small amounts of gas which evolve during periods of nonoperation. The period required for this will vary from tube to tube and from transmitter to transmitter, and is usually a function of the time the magnetron has been inoperative. It is further recommended that this conditioning be carried to levels slightly above but not more than 10 percent beyond the normal operating point of the transmitter to obtain optimum stability.

1. Magnetron tuning drive. The magnetron tuning drive varies the position of the magnetron tuning plugs by use of a reversible motor, mechanical switches, and associated switching and metering circuits. Portions of the magnetron tuning drive circuits are located in the r-f coupler, and the remainder are located in the acquisition receiver control. The spring-loaded FREQUENCY switch, S4, located on the r-f coupler meter panel, may be used to raise and lower the magnetron frequency. A brass tuning dial, graduated from 0 to 100, indicates relative magnetron frequency. The higher the magnetron frequency, the lower the reading on the tuning drive dial. The spring-loaded FREQUENCY switch, S2, on the acquisition receiver control may be used to vary the acquisition magnetron frequency from the tracking console. The FREQUENCY meter at the acquisition receiver control is graduated from 0 to 100 and indicates the relative magnetron frequency. As the magnetron frequency is raised, the meter reading increases. When controlling the motor operation with switch S2, motor X (phase C to neutral) drives the motor. When controlling the motor with switch S4, voltage (phase A to neutral) drives the motor. Motor B2 is a split-phase motor. Capacitors C1 and C5 cause the voltage to one of the windings to be advanced almost 90 degrees. The voltage to the other winding is not advanced in phase. Thus the motor has the proper phases applied to it. The direction of throw of either S2 or S4 determines which winding of B2 will have the advanced phase applied to it. Reversing the phase relationship between the windings reverses the direction of rotation of the motor. FREQUENCY meter M1, located on the front panel of the acquisition receiver control, is calibrated to give an approximate indication of frequency over the entire tuning range of the magnetron. This d-c meter is graduated uniformly from 0 to 100, and a reading of zero should indicate a frequency of 3,100 megacycles per second. Similarly, a reading of 100 should indicate a frequency of 3,500 megacycles per second. Potentiometer R1 on the tuning drive is connected as part of a voltage divider consisting of parallel resistors R2 and R3 connected in series with R1. The network is connected from ground to plus 150 volts. The potentiometer arm of R1 is connected through 10,000 ohms of resistance to meter M1 on the acquisition receiver control and then to ground. When motor B2 turns, it also moves the potentiometer arm of R1. The tapped voltage gives an indication of FREQUENCY meter M1 which varies with the magnetron frequency. As the magnetron frequency is increased, the meter reading also increases. However, the reading on the magnetron tuning head will decrease.

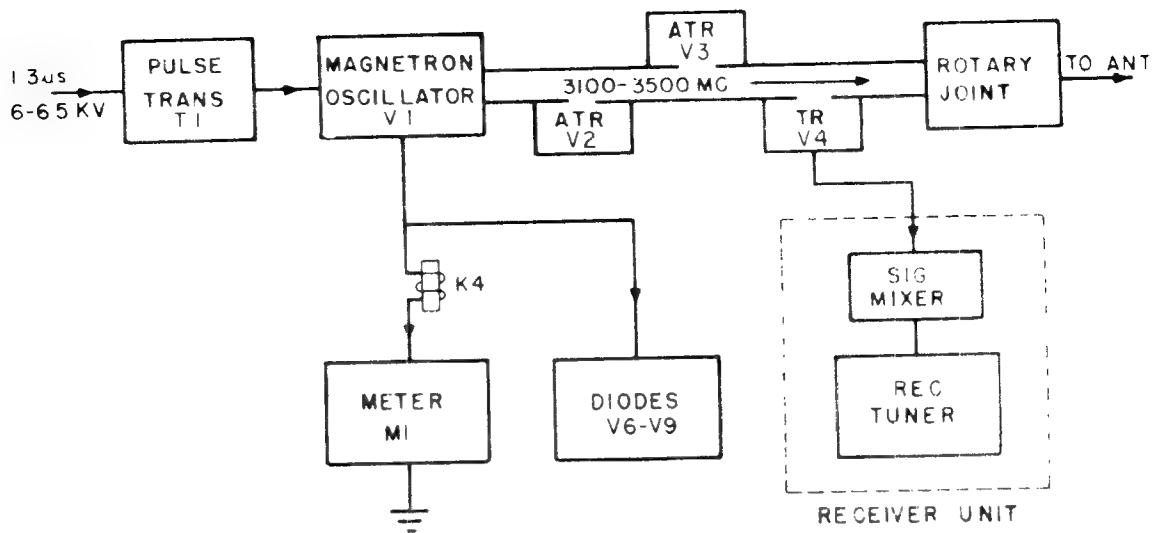
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Figure 8. Acquisition r-f coupler, block diagram.

34. R-F SYSTEM (TM 9-5000-26 pages 196, 197)

a. General. The acquisition r-f system (fig 8) includes all components from the magnetron through the antenna. It includes the arc suppressor, the AFC sampler, the duplexer, the directional coupler, the rotary joint, and the antenna and its controls. The purpose of the r-f system is to carry the r-f energy from the magnetron to space.

b. Arc suppressor. The arc suppressor (TM 9-5000-26, page 195) is a device in the waveguide near the magnetron which prevents arcs from traveling down the waveguide and striking the output window of the magnetron. An arc across the glass window could cause the window to overheat and crack. The action of the arc suppressor is to turn off the transmitter after arcing has been detected in the waveguide at the magnetron, thus extinguishing the arc. The arc suppressor circuit consists of a wire connected across the wide dimension of the waveguide and insulated from it. This wire is connected through relay K5 to 150 volts. If an arc occurs, K5 will be energized and will remove the sync pulse from the acquisition trigger generator so that the transmitter is shut off as long as the arc exists. To keep relay K5 energized for a short period, capacitor C14 is placed in parallel with the coil of K5. Capacitor C15 prevents r-f energy from leaking out of the waveguide at this junction by bypassing it to ground.

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c. AFC sampler. The automatic frequency control requires a sample output of the magnetron frequency for use in its mixer. This output is obtained by inserting an adjustable coupling loop into the magnetic field of the waveguide near the magnetron. This coupling is adjusted for just enough output to make the AFC produce a satisfactory limiter current.

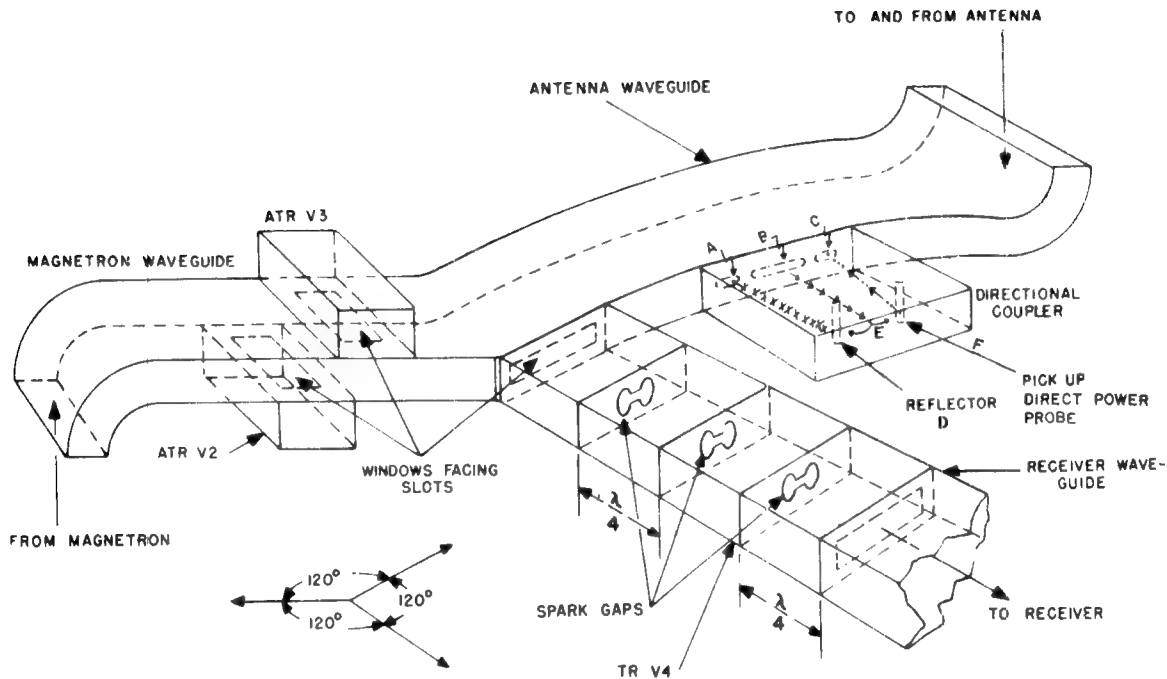


Figure 9. Duplexer and waveguide.

d. Duplexer (fig 9). The use of a common antenna for transmitting and receiving requires a fast acting switching device to disconnect the receiving system from the antenna during the transmitted pulse and to disconnect the magnetron during the remaining period. This switching device employs a transmit-receive (TR) tube, together with two ATR tubes. The duplexer is that portion of the waveguide system which includes the Y-junction and the TR and ATR tubes. It makes possible the use of a single antenna system for both transmitting and receiving. ATR and TR tubes are resonant cavities whose resonant frequency corresponds to the transmitter frequency. They are filled with gas at a low pressure. Spark gaps shorten the time required for ionization and provide maximum protection for the receiver in the TR tube. During the transmitting cycle, the r-f energy traveling down the waveguide sees an open circuit at the ATR tubes, but the transmitted power enters the resonant cavities causing oscillation. The power in the cavities is high because of the energy contained in the transmitted pulse and causes the gas in the cavities to ionize. This reflects a short circuit

across the ATR cavities, and the r-f energy continues down the waveguide to the Y-junction. At the Y-junction the r-f energy sees two paths. Some of the energy enters the TR box. The TR box is made up of four cavities, each a quarter wavelength. The spark gap nearest the receiver has a keep-alive voltage of -800 volts applied to it. Because it has the keep-alive voltage applied to it, this gap requires much less energy to cause arcing and will break down first. In very rapid succession, the second and third spark gaps and the resonant window to the waveguide fire. The transmitted pulse then sees a continuous waveguide and travels toward the antenna. During the receiving cycle, received signals are focused by the antenna into the feedhorn and travel down the waveguide to the Y-junction. These signals do not have enough energy to ionize the TR or ATR tubes. The deionized ATR tubes present a high impedance and prevent the received signal from being dissipated in the transmitter. Since the Y-junction is one-quarter wavelength from the ATR tubes, a low impedance appears at that point. A low impedance at the Y-junction indicates a point of high current and, therefore, high magnetic-field strength. Because the TR tube is coupled into the H (magnetic) field of the waveguide at this point, a maximum value of received signal strength is delivered to the signal mixer. The Nike I ACQ radar uses 2 ATR tubes to insure broad-band operation. These cavities are tuned to frequencies near the upper and lower extremes of the magnetron bandwidth and are so placed on the waveguide that their combined effect over the entire magnetron frequency range is the equivalent of placing one ATR tube one-quarter wavelength from the Y-junction. The TR box, containing three spark gaps, provides protection to the receiver from the power contained in the transmitted pulse of energy. The TR tube may be tested by measuring the voltage at its terminal. The voltage should be 670 ± 50 volts with the transmitter off and 200 volts with the transmitter on. The current through the TR tube should be 130-170 microamperes. The voltage is measured from the keep-alive voltage probe to ground with a TS-352/U multi-meter on the 1,000-volt, 20,000-ohm per volt scale. The current is measured with the TS-352/U multimeter on the 0-1 milliamperes scale with the meter connected between the keep-alive probe and the keep-alive connection on the TR tube.

e. Directional coupler. The acquisition directional coupler is a bi-directional coupler that samples the r-f output and makes this sample available for the measurement of frequency power and standing-wave ratio. It also provides a means of injecting a signal into the waveguide for measurement of receiver sensitivity. It is a straight, inclosed section or waveguide attached to the narrow dimension of the r-f line. Energy is fed into the directional coupler by means of three coupling irises. Two pickup points are located within the coupler. The point marked PICKUP is used in the measurement of direct power, and the point marked REFLECTOR in the measurement of reflected power. In measuring direct power, a pickup probe is inserted at the pickup point and a probe made of a power-absorbing material is placed at the reflector point. The absorbing material is used to dissipate unwanted energy in order to prevent erroneous indications on the power-measuring device. When measuring reflected power,

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the positions of the two probes are reversed. The coupling irises, A, B, and C, are spaced a quarter wavelength apart, and their coupling coefficients are such that the sum of the energy coupled into the inclosed section by irises A and C is equal to the energy coupled in by iris B. The energy coupled in at iris B divides evenly at point E, with half of the energy traveling to point D and the other half to point F. All the energy entering the coupler is attenuated. The attenuation at the various frequencies is stamped on each individual coupler. In the measurement of direct power, energy entering the coupler through A travels a quarter wavelength (90 electrical degrees) to point D. Energy entering at point B must travel the distance ABED. This distance is equal to $3/4$ wavelengths (270 electrical degrees). Therefore, at point D the energy traveling the distance AD is always 180 electrical degrees out of phase with the energy traveling the distance ABED. Since these voltages are equal in amplitude, they cancel. Energy entering at iris C must travel the distance ABCF. This distance equals a $3/4$ wavelength, or 270 electrical degrees. The energy entering at iris B travels the distance ABEF. Since this distance is equal to the distance ABCF, the energy arriving at the direct power probe from both sources is in phase. Therefore, the energy available to the direct power probe at F is directly proportional to the transmitted r-f energy. The directional coupler provides a means of measuring the standing-wave ratio, since this ratio may be computed from the direct power and reflected power measurements. A signal from a signal generator can be injected into the waveguide at point F. Such a signal can be used as a means of measuring the sensitivity of the receiver system. This measurement is made by ordnance personnel.

f. Rotary joint. In order to transfer r-f energy from a fixed transmitter to a rotating antenna, a rotating waveguide connection (fig 13) is used. Because of the nature of the modes and field in a waveguide, it is not possible to merely put a rotating coupler in a waveguide system. Instead, it is necessary to change from waveguide to coaxial cable and let the coaxial section rotate. The fields in coaxial cable are such that this is permissible. In the Nike I acquisition rotary joint, the lower section of the joint acts as a receiving antenna. The rigid center conductor picks up the energy coming toward it from the transmitter and conducts this energy upward, where it is radiated from the knob junction back into the waveguide. By this means the upper half of the joint may rotate without any electrical contact through the center of the coaxial cable. The outside shell of the coaxial section presents another problem. There must be a definite mechanical break in order to allow rotation of the upper section of the waveguide. By using two choke joints at the coupling point, the system reflects a short circuit across the mechanical break. These choke joints are effectively a half-wave shorted section and reflect zero impedance at the coupling. To match the waveguide to the coaxial, the waveguide is tapered. To match impedance at the output of the knob junction, four ridges are used. These ridges may be moved along the waveguide to obtain lowest standing-wave ratio. This is a factory adjustment which

is a compromise for the band of frequencies used and should not be disturbed by the repairman. A combination of a window and choke coupling are used at both ends of the rotary joint assembly to allow pressurization of the joint. When a full-power magnetron is used, there is a tendency for the rotary joint to arc because the waveguide going into the rotary joint is tapered, resulting in the spacing of components in the rotary joint being closer together than elsewhere. To eliminate this arcing, a pressurized rotary joint is used. Just above the choke joint between the r-f coupler unit and the antenna drive unit, the waveguide contains a window which passes the r-f energy but blocks air. There is another such window just above the rotary joint. The joint itself is pressurized by a unit located near the center of the antenna drive unit. The unit is accessible by swinging out the resolver amplifier chassis. It maintains a pressure of 10 to 16 pounds per square inch inside the rotary joint. The pressurization unit starts when radar power is turned on but may be turned off by a switch on the unit itself if the system will work satisfactorily without pressurization. The pump is powered by a 208-volt, 3-phase, 400-cycle, 1/45-horsepower motor. The pump has a capacity of 110 cubic inches of air per minute at a pressure of 10 to 16 pounds per square inch. A pressure switch controls a relay on the motor primary so that the unit starts when the pressure falls to 10 pounds and shuts off when the pressure reaches 16 pounds as read on the pressure gage. A safety valve is included which will prevent the pressure from exceeding 30 pounds per square inch. This unit is designed for intermittent operation with a duty cycle of 1 minute on and 2 minutes off. The rotary joint is not airtight so some leakage of air is to be expected. If the motor runs more frequently or for longer periods of time, the rotary joint is leaking air excessively and should be repaired or replaced.

g. Waveguide. The waveguide used in this system is for the 10-centimeter band. It is a rectangular guide 1 1/2 inches by 3 inches silver plated on the inside to reduce losses. A waveguide will pass all frequencies higher than the frequency at which a half wavelength is equal to the width of the guide. In Nike I, the guide must be greater than 5 centimeters wide. This guide is 8 centimeters wide and will pass all frequencies in the 10-centimeter band. The wide dimension of a waveguide may not be made too wide compared with a half wavelength, or undesirable modes of operation result. The narrow dimension of the guide is not critical except that it must be great enough to prevent arcing. The maximum voltage in a waveguide exists between the centers of the two wide faces, and therefore, the narrow dimension must be great enough to withstand the maximum voltage expected. Standing waves in a waveguide result from impedance mismatches, usually at the two ends of the line. However, any discontinuities in the line tend to create standing waves. Some components in this system which cause standing waves are: the coupling from the magnetron, coupling in and out of the rotary joint, the feedhorn from waveguide into air, and finally nearby objects which affect the loading of the antenna. A measure of the relative quality of a

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waveguide system is called its standing-wave ratio or SWR. The comparison is 1:1 for a perfect line, 1.2 to 1.6:1 for an average line and may go to 25.1 for a line in need of repair. The directional coupler is used to determine this ratio. The best method of evaluating this information is to compare the present SWR as measured with the ME-51 power and frequency meter with previous measurements. An unusual change would indicate repairs are needed.

CHAPTER 4

ACQUISITION ANTENNA SYSTEM

Section I. DESCRIPTIVE DATA

35. ANTENNA AND ELEMENTS (TM 9-5000-26 page 205)

The acquisition antenna is of the multiple-reflection type. Three reflections are used to focus the radiated beam. These reflecting surfaces produce a beam 1.5° wide and 5.5° high. By changing the curvature and angle of the reflector, the beam can be changed in shape and can be directed at an angle of from 2° to 22° above the horizontal. The antenna consists of five main elements (fig 10):

- a. Feedhorn for matching impedance between the waveguide and the air.
- b. A pillbox radiator which consists of a parabolic chamber used to shape the beam horizontally.

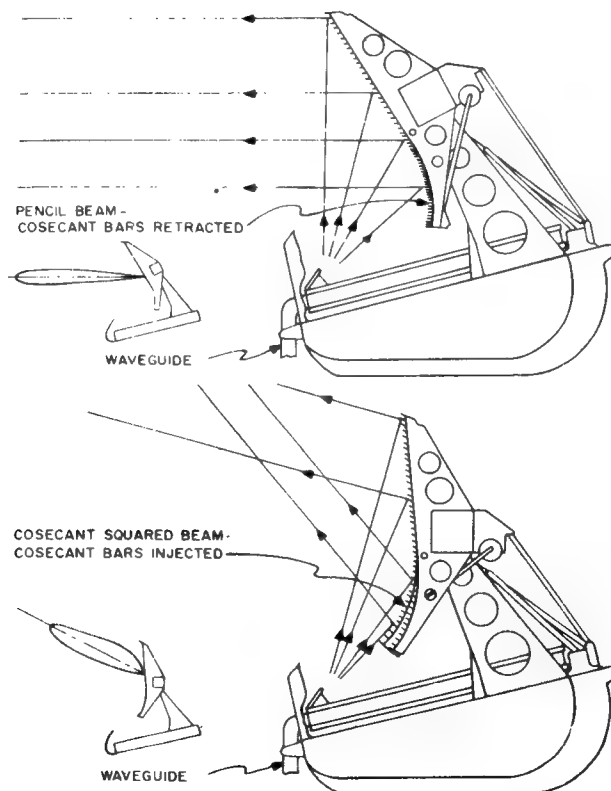


Figure 10. Reflector bar assembly.

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c. A main reflector grating which permits the radiated beam to be variable in elevation and shapes the beam vertically.

d. Cosecant bars which may be inserted between the bar grating of the main reflector to change its curvature. These are called cosecant bars because of the shape of the radiation pattern that results when they are injected.

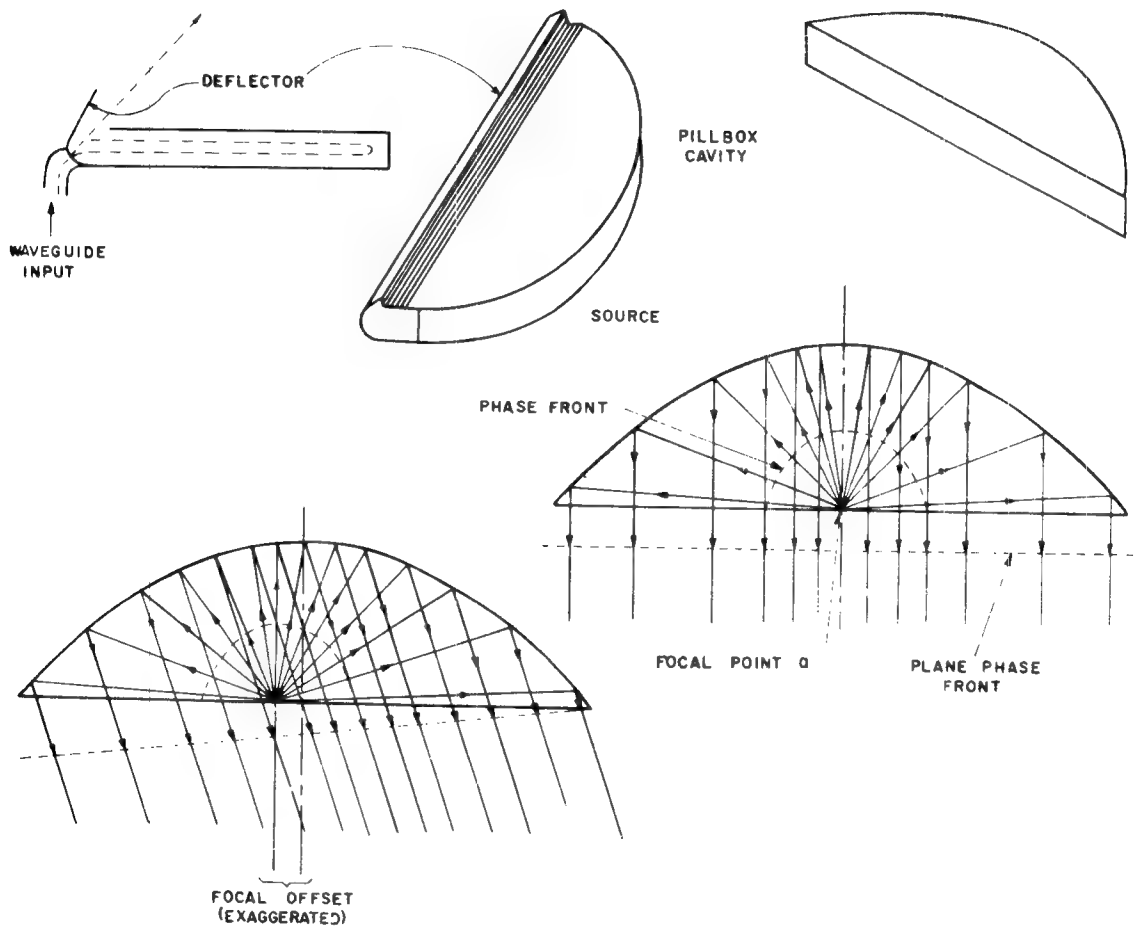


Figure 11. Acquisition antenna pillbox.

36. DESCRIPTION OF REFLECTOR

The reflector elements are housed in a Fiberglas radome 5 feet by 5 feet by 15 feet. This radome keeps dust and water from entering the antenna. It has no adverse effect on the antenna radiation pattern. The main reflector may be tilted through 9° by a hydraulic system. This hydraulic system also can be adjusted to insert the cosecant bars at any desired elevation, thereby changing the elevation pattern from a 5.5° beam to a large, cosecant-squared beam. The

main reflector being tilted causes the radiated beam to be adjustable in elevation from 2° to 22° from the horizontal. This elevation can be set electrically from the van either to a fixed value, or it can be scanned continuously, depending on the tactical situation. This flexibility enables detection of targets to 60,000 feet in altitude and 120,000 yards in range. The antenna may be turned in azimuth at a speed of 10, 20, or 30 revolutions per minute, depending on the range and speed of the target. Radiation from the antenna takes place as a result of several reflections. Energy from the feedhorn is radiated into the pillbox radiator (fig 11). There it strikes the parabolic wall where it is focused into parallel rays and a straight phase front. These parallel rays strike the deflector and then the main reflector. Radiation from the antenna is offset 1° in azimuth as a result of a slight offset of the feedhorn from the focal point of the parabola to reduce standing waves in the pillbox radiator.

37. CONTROLS

a. Coverage and scan. The elevation coverage and the rate of azimuth scan are both controllable from the acquisition control panel located on the battery control. When the elevation coverage switch is placed in the up position, the beam angle tilts up to 22° maximum. When the switch is in the down position, automatic scan takes place between 2° at the lower limit and a preset upper limit depending on the scanning mode desired.

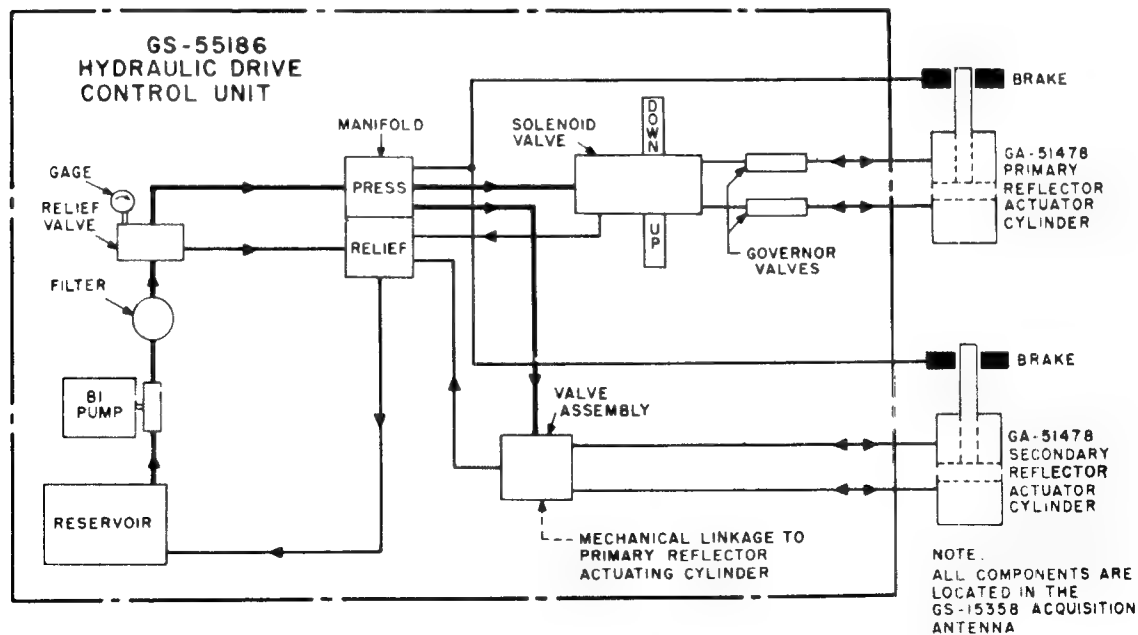


Figure 12. Antenna hydraulic system.

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b. Hydraulic system (TM 9-5000-26, page 204). The hydraulic system for controlling elevation coverage (fig 12) is located under the rear side of the radome. This hydraulic system consists of an oil pump, filter, valves, pistons, and a reservoir. The pump delivers 10 to 20 pounds per square inch (psi) as indicated on the pressure gage as soon as ACQ POWER is turned on. However, when a control switch is operated, the pressure increases to 250 psi, and the valves

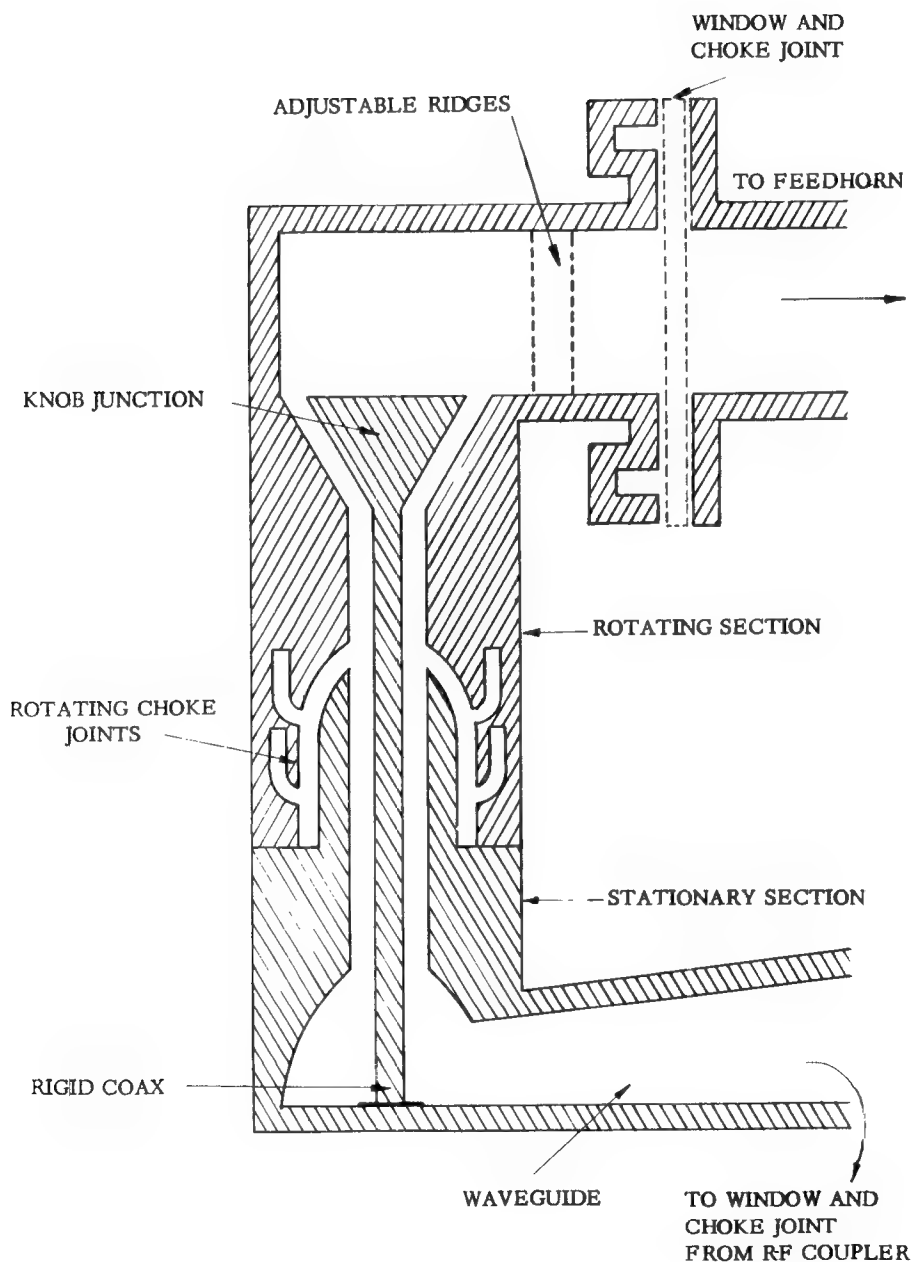


Figure 13. Cutaway view of rotary joint.

direct the oil to the desired pistons, causing the reflector to tilt. This pump is capable of delivering 3,000 psi. Its output is regulated at 1.2 gallons per minute so that the reflector tilts at a uniform rate. A unique feature of this system is that when no valves are operated, the pressure is released and returned to the reservoirs. Pipe resistance causes the meter to read 10 to 20 psi when the reflector is not being positioned. As long as the pressure is below 80 psi, two hydraulic brakes prevent either the main reflector or the cosecant bars from moving. The two valves which control the tilt of the reflector are the UP and DOWN valves. When the reflector is being elevated, a mechanical linkage from the top of the reflector to a sliding vertical shaft causes an INJECT valve to open at a preset angle. This causes the cosecant bars to be driven forward. When the reflector is depressed, the sliding shaft moves upward, closing the INJECT valve and opening the WITHDRAW valve. This action causes the cosecant bars to be withdrawn. In automatic operation the reflector is successively elevated and depressed within angular limits determined by the position of two limit microswitches. Ground may be applied to either the UP or DOWN solenoid winding through contacts on the 3/6 position rotary acquisition SCAN-AZIMUTH switch, S6, at the acquisition control panel. This arrangement insures that control of elevation scan is possible from the console, which also controls the rate of azimuth scan. Assume that S6 is in the UP position (contact 2 made to contact 1). Ground is then applied to the UP solenoid, completing the circuit to the minus 28-volt supply through the coil of the UP solenoid. The UP valve opens, applying hydraulic power to the elevating mechanism. The reflector is then tilted upward. During its upward movement, at an angle dependent upon the setting of the mode of scan controls, the sliding shaft causes the INJECT valve to open. Hydraulic power is then applied to the cosecant bar injecting mechanism, and the cosecant bars are driven forward. The entire injection operation is completed during the time the tilt of the reflector changes one degree. When the reflector reaches its maximum 9° angle of tilt, a mechanical stop prevents additional tilt, and hydraulic power is removed through a pressure release valve. The reflector will remain at this maximum angle until the controlling switch is placed in the DOWN position. When the 3-position toggle switch is in the DOWN position, the circuit is completed to the DOWN solenoid through contacts 2 and 3 of K1, and the DOWN valve is opened. Hydraulic power is applied to the depressing mechanism, causing the reflector to be tilted in a downward direction. As the reflector tilts downward, the sliding vertical shaft moves upward. At the lower limit of reflector tilt, a disk mounted on the shaft closes microswitch S2, which is normally open. Relay K1 becomes energized as S2 closes. When K1 is energized, the DOWN solenoid is deenergized (contacts 2 and 3 of K1 open), and the UP solenoid is energized (contacts 1 and 3 of K1 close). In addition, contacts 4 and 6 provide a holding circuit for the coil of K1 through upper limit switch S1, which is normally closed. The reflector is then elevated until the disk on the vertical shaft opens the upper limit microswitch. This deenergizes relay K1, causing the UP valve

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to be closed and the DOWN valve to be opened. This automatic scan cycle continues until the controlling switch is operated out of the DOWN position. The setting of lower limit switch S2 cannot be changed. The setting of upper limit switch S1 and the angle of cosecant bar injection are both determined by which scan mode is set up.

Section II. SCANNING MODES AND COVERAGE

38. SCANNING MODES (TM 9-5000-26, page 205)

a. Mode determination. Two facts determine what mode of scanning is being used. The first is the elevation angle at which the cosecant bars are inserted and

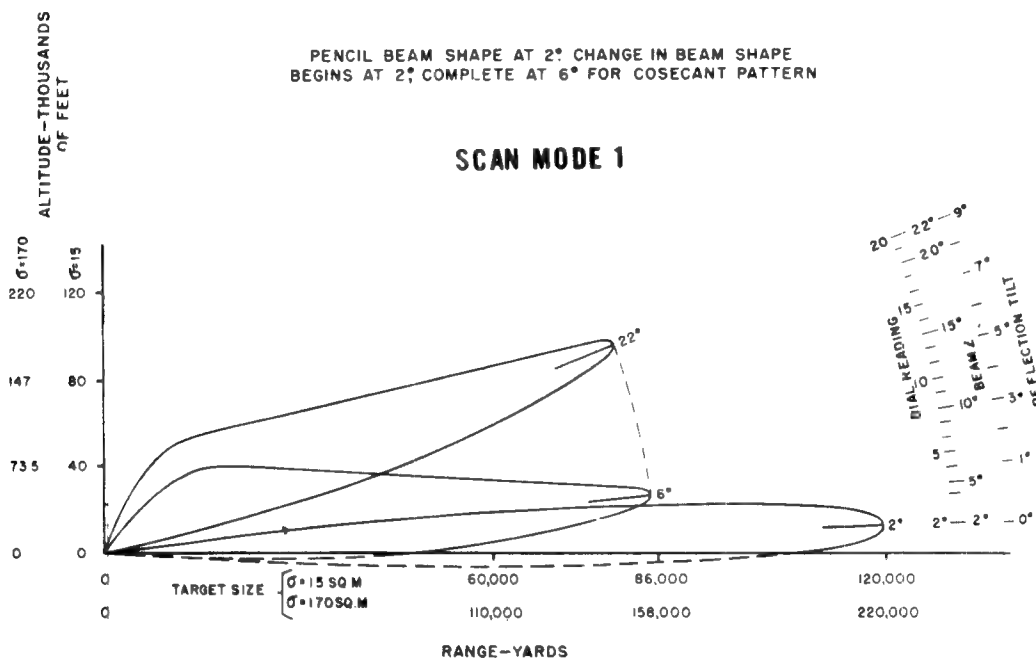


Figure 14. Scan mode 1.

the second is the maximum elevation angle for the radiated beam. The mode used depends upon the set location and the expected altitude of enemy aircraft. Table I shows the data for each mode. The factory initially sets the antenna in mode number 2 (fig 15). These values are continuously variable, and the four modes are only recommendations. Beam coverage patterns for the other three modes are shown on figures 14, 16, and 17. The hydraulic system can tilt the reflector between 0° and 9°. The acquisition elevation coverage dial at the acquisition control panel indicates the elevation of the r-f beam in degrees. The scale is graduated in 2° increments from 2° to 22°. The corresponding angles of reflector tilt are zero and 9°, respectively. At the minimum angle of reflector tilt, 0°,

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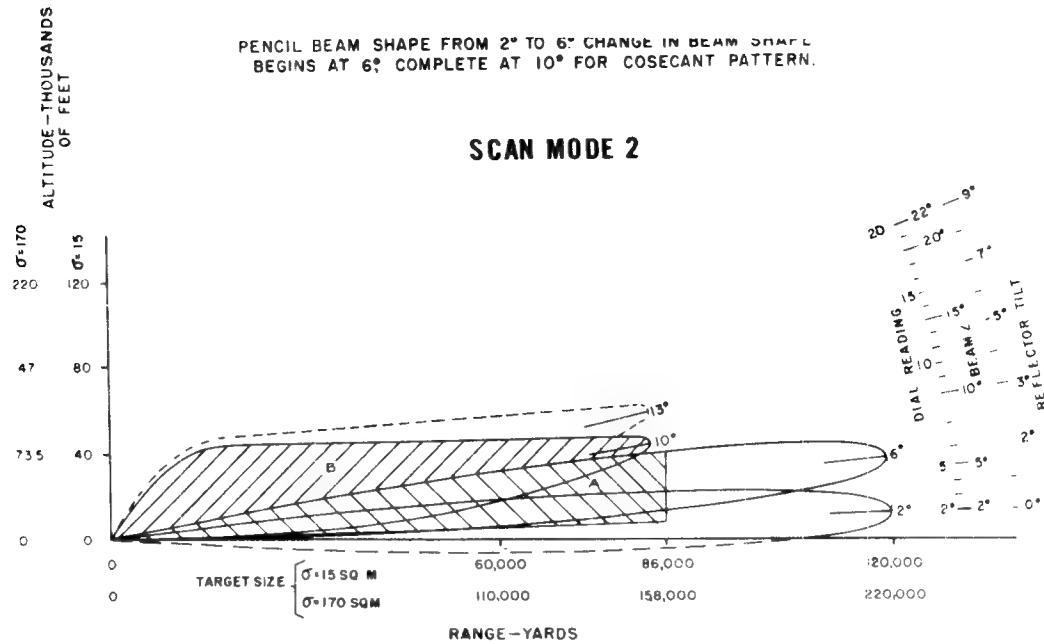


Figure 15. Scan mode 2.

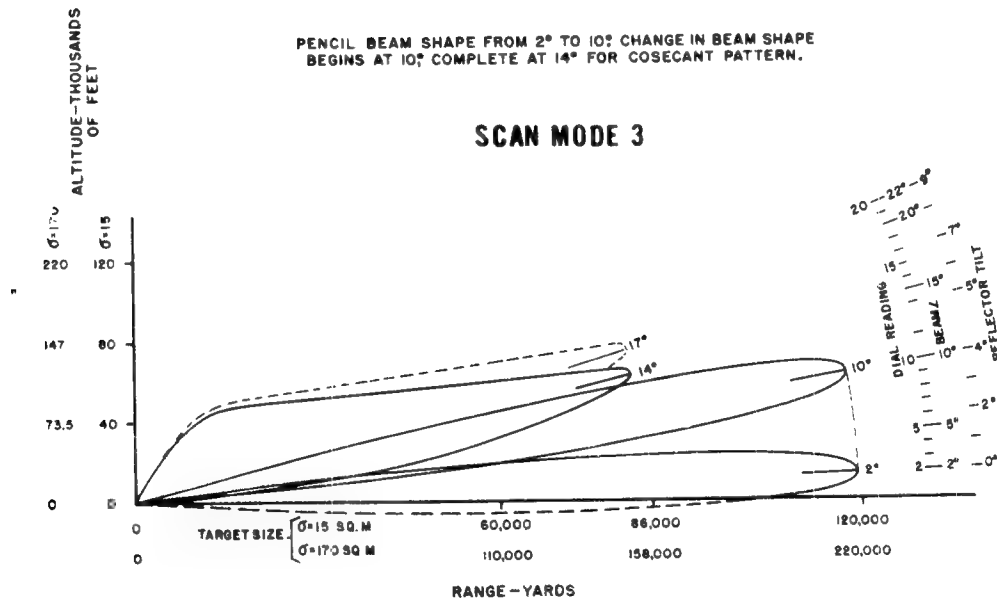


Figure 16. Scan mode 3.

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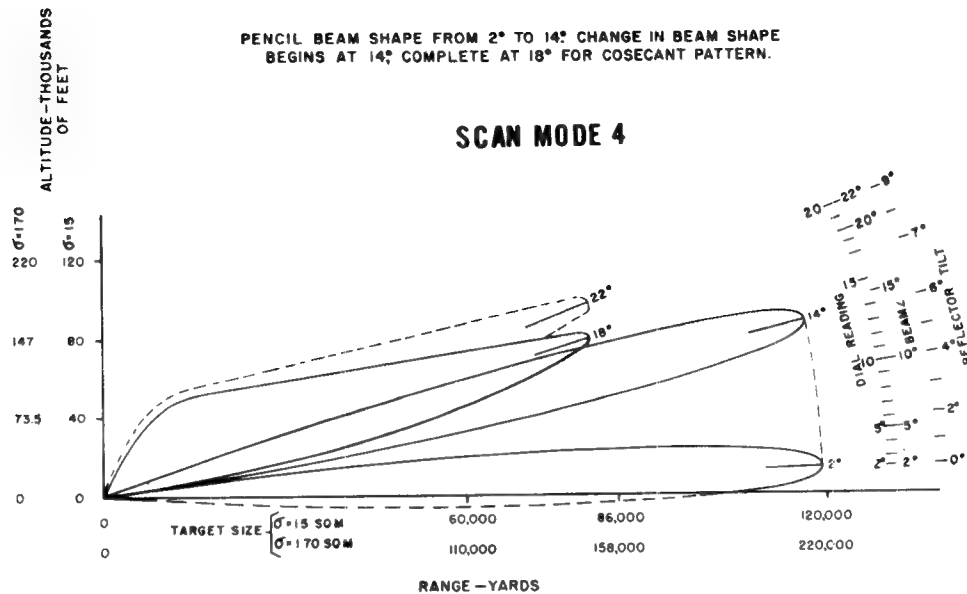


Figure 17. Scan mode 4.

the axis of the r-f beam is 2° above the horizontal. Each degree of reflector tilt causes the beam to be elevated two degrees. The injection of the cosecant bars (which transform the beam from its needle shape to a cosecant-squared shape) causes the beam angle to increase 2° additionally. The injection action is completed during 1° of reflector tilt. The foregoing data can be summarized by saying that during 1° of reflector tilt the cosecant bars are injected.

Table I. Antenna scan mode data.

Scan Mode	PENCIL BEAM		COSECANT BAR INJECTION BEGINS		COSECANT BAR INJECTION COMPLETED		UPPER SCAN LIMIT		Complete Scan Period in Seconds
	Re-flector tilt	Beam Angle	Re-flector tilt	Beam Angle	Re-flector tilt	Beam Angle	Re-flector tilt	Beam Angle	
1	0°	+2°	0°	+2°	+1°	+6°	+9°	+22°	40
2	0° TO +2°	+2° TO +6°	+2°	+6°	+3°	+10°	+3°	+10°	20
3	0° TO +4°	+2° TO +10°	+4°	+10°	+5°	+14°	+6.5°	+17°	28
4	0° TO +6°	+2° TO +14°	+6°	+14°	+7°	+18°	+9°	+22°	40

b. Scanning mode adjustment. To change the scanning mode, the following procedure is used.

- (1) Remove the cover from the hydraulic compartment.
- (2) Move the locknut just above the upper limit switch up or down until the pointer behind it indicates the upper scan limit. Relock the nuts above and below the upper limit switch.
- (3) Loosen the locknut between the two disks to the left of the upper limit switch.
- (4) Adjust the reflector tilt (use jogging switch) until the upper disk points to the desired angle for cosecant bar injection. Adjust the space between the disks until the lower disk is just pressing against the valve below it. The position of the lower disk is adjusted by rotating it on its threaded shaft.
- (5) When the adjustments are complete, retighten the locknut between the two disks and replace the hydraulic compartment cover.

c. Elevation coverage synchros (TM 9-5000-26, page 202). The acquisition elevation coverage dial located on the acquisition control panel in the battery control trailer (fig 18) indicates the degree of tilt of the r-f beam. The system uses a transmitting synchro, located in the acquisition antenna assembly and a repeating synchro on the acquisition control panel. The system uses phase C for synchro excitation. Adjustment of the **system** is made at the repeaters by loosening the bolts holding the repeaters and turning the repeaters. Adjustment is made so that when the transmitter reads 0° (on the 0° to 9° scale in the hydraulic compartment), the repeaters will read 2° (0° of antenna tilt equals 2° of r-f beam elevation).

39. AZIMUTH DRIVE AND CONTROLS (TM 9-5000-26, page 203)

a. General. The acquisition antenna may be rotated at a speed of 10, 20, or 30 rpm. This speed is variable because of the variation in range of targets. At long range, slow rotation is desired to get maximum echo signal, while fast rotation is needed for close-range targets due to their fast angular movement. To rotate the antenna at the different speeds, the antenna drive unit contains a 3-speed, 3-phase motor, a gear reduction box, and a series of control switches.

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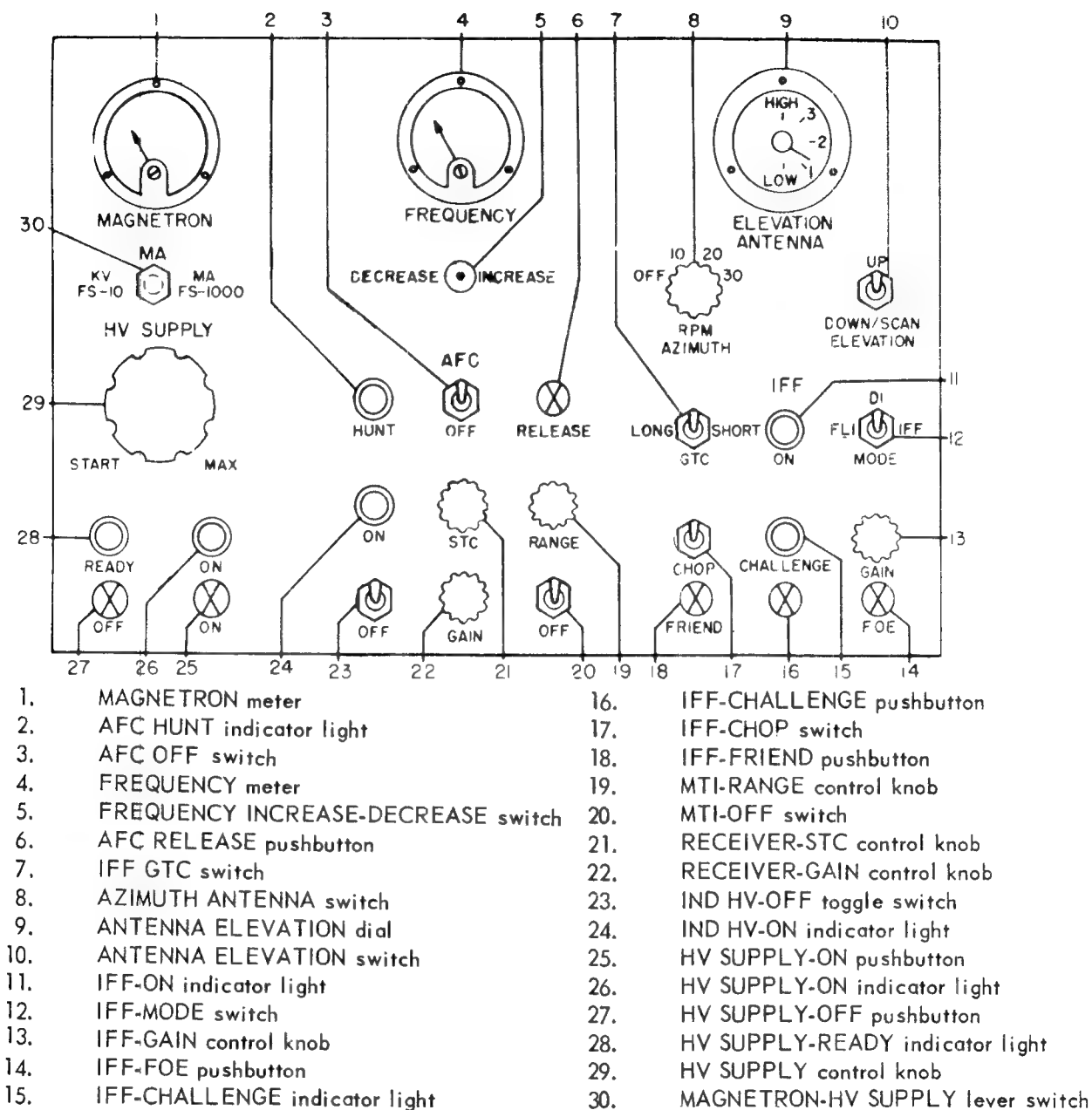


Figure 18. Acquisition control panel.

b. Control switch. The AZIMUTH ANTENNA switch, S7, is a four-section, four-position rotary switch located on the acquisition control panel of the battery control console and is normally operated by the acquisition operator. The four positions are OFF, 10, 20, and 30. When turned to position 10, 20, or 30, the

appropriate relay in the antenna drive contactor is energized. Positions 10, 20, and 30 cause the antenna to rotate at 10, 20, and 30 rmp, respectively. When the switch is in the OFF position, rotation of the acquisition antenna is impossible, and the acquisition indicators are off. This last feature serves to protect the scope.

c. Safety switch S1. This switch is located under a protective cover on the side of the acquisition antenna drive unit. Its contacts are in series with the minus 28-volt supply and are normally closed. When S1 is operated, the energizing potential is removed from the relays of the drive contactor, and the antenna drive unit is immobilized so that work may be performed at the acquisition antenna without danger to personnel.

d. Antenna drive contactor, K1 (TM 9-5000-26, page 203). The antenna drive contactor is located in the acquisition antenna drive. This unit is controlled by switch S7. Three relays and associated wiring make up the contactor. Any one of the three relays may be energized through S7. Three-phase, 208-volt, 400-cycle power is applied to the windings of the drive motor through contacts of the energized relay. The relay energized determines the speed of rotation of the antenna. A varistor is connected across each relay coil. A varistor decreases resistance with an increase in the applied voltage. This characteristic eliminates the surge voltage which would otherwise appear when the circuit of the relay coil is opened. In the antenna drive contactor, the inductance of the three relay coils is quite large. Removal of the -28-volt energizing potential will cause a pulse voltage of considerable amplitude to appear across the coil. In the absence of varistors, this voltage would cause objectionable arcing across the contacts in series with the coil. The varistors across the coils provide an effective short circuit for any pulse of voltage and thereby prevent arcing across switch or relay contacts.

e. Drive motor. The drive motor (fig 19) is located beneath the turntable and gear reduction box in the antenna drive. It is a 3-phase, 3-speed, 400-cycle, constant-torque, 4-horsepower, a-c induction motor of the squirrel-cage type. Characteristics of operation of the motor are shown in table II. The drive motor is connected to the gear reduction box by means of a slip clutch, provided to protect the motor in the event the mechanism should become jammed. In such a case, the motor would continue to rotate, but the clutch would slip. This slip-page would prevent damage to the motor. The gear reduction box contains gears which provide a 30 to 1 step-down ratio. It is oil filled and sealed to prevent the entry of dust and to reduce the noise level. With the motor turning at 3,600 rpm, the output shaft of the gear reduction box will turn at 120 rpm. The output shaft of the gear reduction box is keyed to a 14-tooth drive gear. This gear drives an idler gear which drives a large gear below the antenna, and the antenna is rotated

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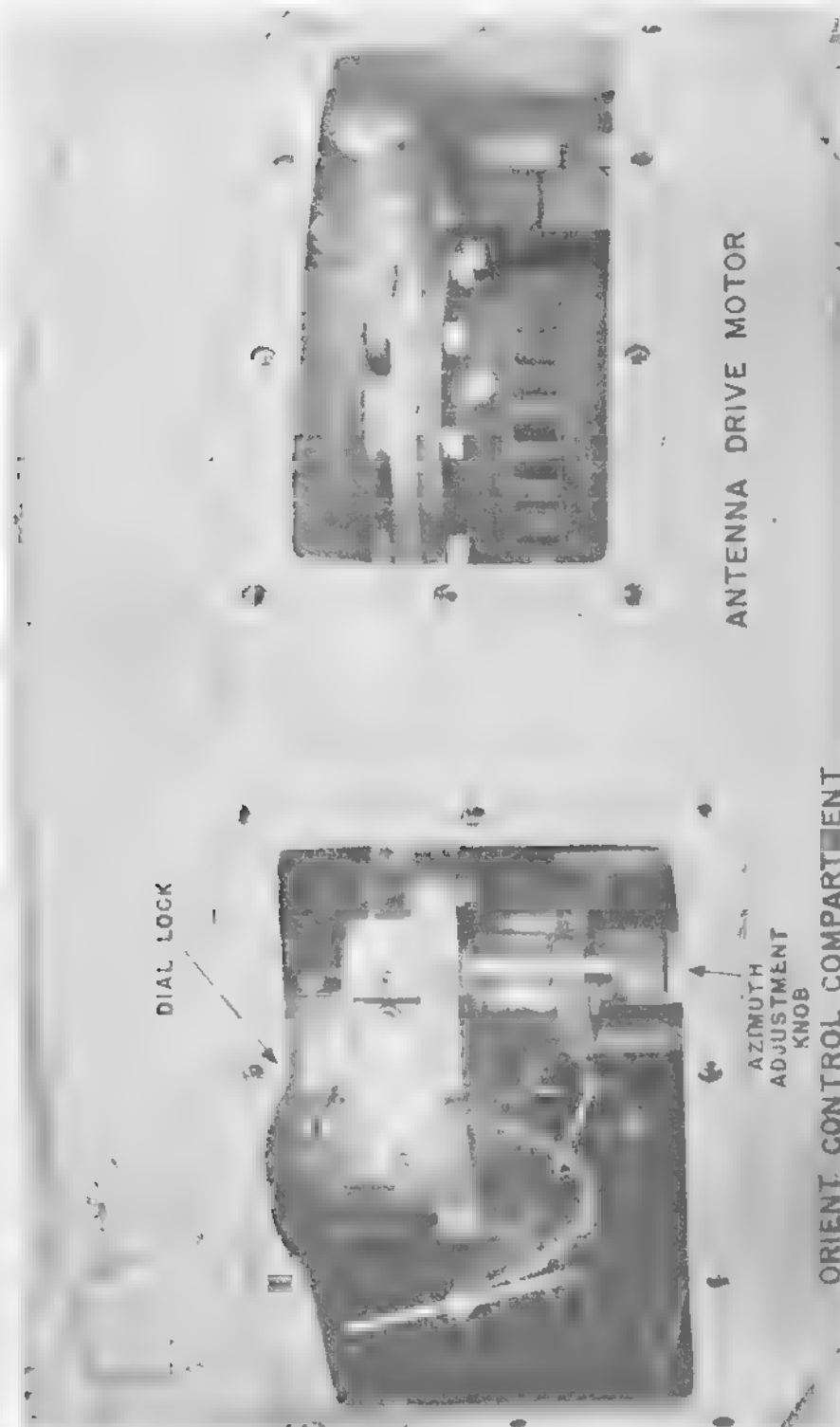


Figure 19. Acquisition antenna drive.

Table II. Drive motor characteristics.

ANTENNA SPEED	MOTOR SPEED	HORSE- POWER	CURRENT FLOW
10 rpm	3,600 rpm	0.75	5.7 amps
20 rpm	7,200 rpm	1.5	7.2 amps
30 rpm	10,800 rpm	4.0	18.3 amps

at 10 rpm. The over-all reduction of the drive gear, idler, and main antenna gear is 12 to 1. With a motor speed of 3,600 rpm, the output shaft of the gear reduction box will turn at 120 rpm, and the turntable itself will rotate at 10 rpm. A 230-tooth gear is mounted beneath the main gear. This gear is meshed with another gear identical to it, which provides rotation of the acquisition azimuth resolver and the azimuth dial. The turntable moves on 6 bearings which are fastened to the side of the drive unit. With each of these 6 bearings is an adjustable companion bearing located above the turntable. The companion bearings insure smooth rotation of the turntable to maintain proper tension on the bearing surfaces. A cover plate is mounted on the turntable. A lip on the cover plate fits a groove around the top of the drive unit. The groove is filled with antifreeze, and the lip is immersed in this liquid to prevent the entry of dust. Failure to keep the groove filled with liquid will result in damage to the mechanism caused by foreign substances entering the unit. The operation of the azimuth drive with the rotary switch(S7) on the acquisition control panel in the 10-rpm position (position 2) is as follows:

- (1) Minus 28 volts energizes the 10-rpm relay in the azimuth contactor drive. The current flows through the safety switch, S1, through the 10-rpm relay through the position 2 contacts of S7, and to ground.
- (2) When the 10-rpm relay is energized, it will close its contacts so as to apply 3-phase, 208-volt power to the T1, T2, and T3 windings of the motor. With power applied to these windings, the motor will turn at approximately 3,600 rpm and drive the antenna at approximately 10 rpm.
- (3) The 20-rpm relay when energized will connect power to windings T11, T12, and T13, and the 30-rpm relay connects power to T21, T22, and T23 of the drive motor. The windings energized determine the number of poles energized in the field of the drive motor, which determines the speed at which the motor will be driven.

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- (4) The drive motor contactor is so designed that only one relay can supply power to the motor at one time. From the schematic, it can be seen that if the 10-rpm relay, having been energized, sticks and will not release, it will hold contacts in the energizing circuits of the 20- and 30-rpm relays open, so as to prevent more than one set of poles in the drive motor from being energized at any one time.

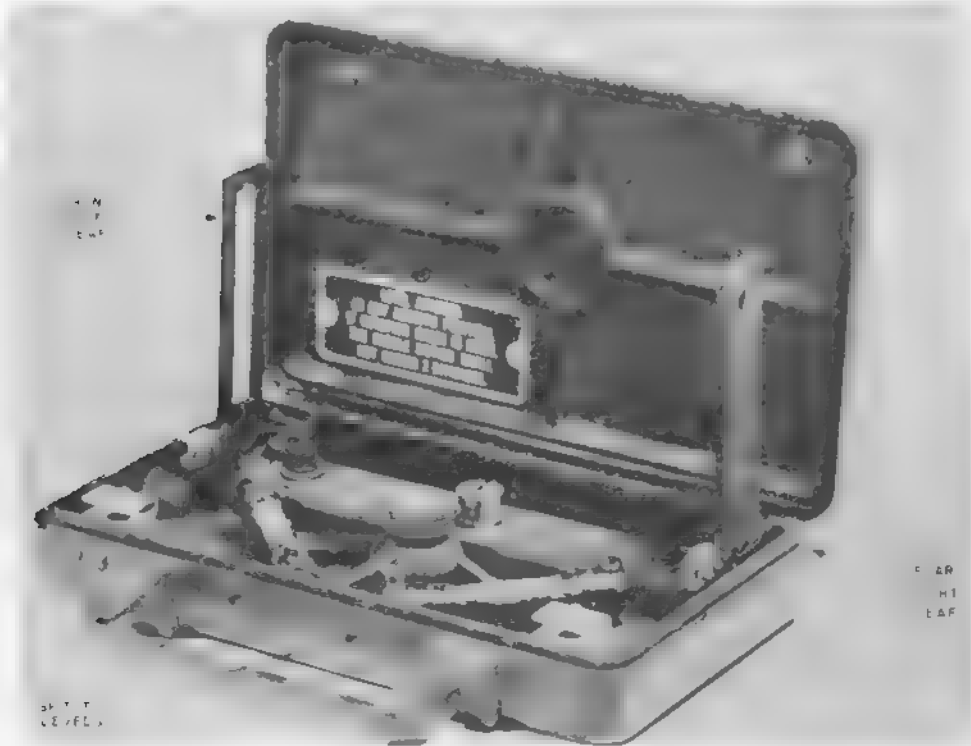


Figure 20. Acquisition orientation test set.

40. ACQUISITION ORIENTATION TEST SET

The acquisition orientation test set (fig 20) is used in the emplacement of the equipment to level the acquisition antenna and to make certain that its direction is the same as that of the track antennas. During emplacement, the level box is placed on the rotating platform of the antenna set upon two dowel pins, and the tripod part-leveling adjustments are set so that the spirit-levels on the peep sight and level box are level over the full 360° travel. When the leveling has been completed, the plane table portion of the peep sight is opened up, and a suitable target in the surrounding area is selected. The antenna is rotated until the vertical cross hair splits the target. The track antenna is then aimed so that the chosen target is centered, using the track optics for sighting purposes. Indications on the oscilloscope indicators are then checked to verify the alinement.

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Adjustments to insure that both systems are pointing in the same direction are made in the acquisition transmitter section by moving the outside container or case of the antenna resolver of the acquisition system. Care must be taken to make certain that only the peep sight and level box assigned to the antenna is used in sighting it, because the dowel pins on the platform have been adjusted to the particular serial number of the level box. If it is necessary to use a different serial number with the equipment, it will be necessary to readjust several screws inside the level box to make certain that the sighting direction is correct. This can be done in conjunction with the indicators and the telescopes of the tracking antennas.

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CHAPTER 5

ACQUISITION RECEIVER SYSTEM

Section I. RECEIVER COMPONENTS

41. INTRODUCTION

The receiver system of the acquisition radar functions to produce from the weakly received r-f energy a video signal of sufficient strength to be presented on the acquisition scopes. Since radar frequencies are in the microwave region and since no practical method has been devised to amplify these frequencies, the r-f energy present at the antenna must be changed to a lower frequency which will allow it to be amplified by conventional means. In order to accomplish this, the radar receiver is of the superheterodyne type. The superheterodyne receiver permits double detection of the signal. It changes the frequency of the returned echo to a lower frequency and then amplifies that lower frequency. The receiver then detects the pulse envelope, further amplifies these pulses, and displays them on the PPI and PI.

42. RECEIVER BLOCK DIAGRAM (TM 9-5000-26, page 207)

a. Antenna and waveguide. The same antenna and waveguide that are used for the transmitter are used for the receiving system. Targets as well as terrain features reflect portions of the transmitted energy to the antenna. The antenna reflector focuses these reflections into the feedhorn. From the feedhorn they travel through the waveguide and rotary joint to the TR tube. The ATR tubes prevent the reflections (r-f energy echoes) from entering the magnetron. Since the energy contained in these echoes is not sufficient to cause arcing of the TR tube, the energy enters and travels through the TR tube to the preselector. The antenna, waveguide, ATR tubes, and TR tube were described in chapter 3 and will not be discussed in this chapter. The student should remember that they are parts of the receiver as well as the transmitter. Faults occurring in these components will affect the efficiency of both the transmitter and the receiver.

b. Preselector. The preselector is a cylindrical resonant cavity tunable over the entire range of the acquisition magnetron and is functionally part of the mixer channel. The transmitter energy which has been reflected enters the cavity from the TR tube. The preselector acts as an r-f amplifier stage in that it allows only the r-f energy which is at the proper frequency to pass through without being unduly attenuated. Unlike an r-f stage, however, it does not amplify the r-f energy. Even the r-f energy at the proper frequency is attenuated slightly. The

cavity is broadbanded, with a bandpass of approximately 25 megacycles. Frequencies outside this bandpass are greatly attenuated. The preselector thus acts to improve the selectivity of the receiver. The r-f energy is coupled out of the preselector into the signal mixer.

c. Signal mixer. The signal mixer is functionally part of the mixer channel. The r-f energy coupled out of the preselector cavity is mixed in the signal mixer with the output energy from the local oscillator, the frequency of which is 60 megacycles higher than that of the received echo. Heterodyning takes place in the mixer and produces a beat frequency of 60 megacycles. This signal is known as the intermediate frequency of the radar and now contains the intelligence originally carried by the received echo. These 60-mc, i-f signals are produced by a crystal detector in the signal mixer. From the signal mixer the signal is fed to the i-f preamplifier via a coaxial cable.

d. Local oscillator. The local oscillator is a reflex klystron capable of producing r-f energy of frequencies 60 megacycles above the transmitter throughout the entire band of transmitter frequencies. The frequency of its output is controlled by the transmitter through the automatic frequency control (AFC) unit. One output of the local oscillator is coupled into the crystal mixer where it is mixed with the received signal to produce the 60-megacycle i-f signal. A second output of the local oscillator is coupled to the AFC mixer, where it is mixed with a sample of the transmitter's output to produce an intermediate frequency which is used to control the local oscillator's frequency.

e. I-F channels. The outputs of the signal mixer are pulses of i-f energy at a frequency of 60 megacycles. The energy level of these pulses is extremely low. These pulses, after detection, are to be displayed on the indicator scopes located in the van. The signal mixer, itself, is located in the acquisition antenna assembly, which is a great distance from the van. Therefore, the first stages of i-f amplification must be at the acquisition r-f coupler so that the weak i-f signals will not be attenuated nor mixed with noise while being transported over the cable. The i-f amplifier channel is located close to the signal mixer. It contains two stages of low-gain (low-noise) amplification and three stages of high-gain amplification. The i-f output of the preamplifier is fed over a long coaxial cable to the i-f attenuator located in the van. The attenuator allows for an attenuation of the i-f signal of from 20 to 40 decibels in steps of 5 decibels. This variable attenuation gives the operator a convenient means of increasing the gain of the receiver to compensate for loss of gain as the elements age. The output of the attenuator receives further amplification in seven more high-gain stages of the i-f amplifier channel. The last stage of the i-f amplifier channel is a detector where the video envelopes of the i-f pulses are detected. This comprises the second detector of the radar receiver.

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f. AFC channel. The acquisition transmitter can be rapidly tuned by the operator over its entire frequency range. In order to take full advantage of this feature, the receiver must be tuned automatically to follow the change in transmitter frequency. This is the function of the AFC channel. The AFC channel receives a sample of the transmitter pulse and a portion of the output of the local oscillator. It mixes these frequencies and amplifies the resulting intermediate frequency. The i-f signal is then fed to a frequency discriminator, which determines whether it is above or below the desired 60 megacycles. The discriminator then tunes the local oscillator through a servo system so that the resulting intermediate frequency becomes 60 megacycles.

g. Sensitivity time control channel. Targets and clutter close to the acquisition radar return very strong echoes to the receivers. This causes the center of the PPI's to be extremely bright. This brightness may prevent the operator from seeing a target at a greater range. The sensitivity time control (STC) reduces the gain of the receiver for close ranges by applying a negative square wave with a trailing slope to the grids of the last three stages of the i-f preamplifier channel. The square wave starts with preknock and extends to about 4,000 yards. The trailing edge may extend to 20,000 yards, if desired. The STC control simply controls the amount of gain reduction at ranges up to 20,000 yards.

h. Moving target indicator channels. Targets located at the same range and azimuth as clutter cannot be detected on the PPI's, and targets close to clutter in range and azimuth are difficult to detect. The moving target indicator (MTI) reduces the brilliance of the clutter display on the PPI's without materially affecting the brilliance of target displays. The MTI compares two target video pulses resulting from two successive bursts of energy from the transmitter. These video pulses are arranged to have opposite polarity and will cancel if they both occur at the same time and are the same amplitude. However, if the target has moved during the time interval between these pulses, the video pulses will occur at different ranges and will not cancel completely. The noncanceled video is amplified and applied to the PPI's. Thus, the MTI distinguishes between moving targets and the stationary targets which produced clutter. In this MTI, large masses of clutter are not completely eliminated from the PPI's, but their brilliance is materially decreased, enabling the operator to detect moving targets in or near them. The MTI does reduce the brilliance of all targets, including moving targets to some extent. This feature, combined with the fact that clutter is not a serious problem at long range, dictates a design feature of the MTI which limits its use to ranges up to 35,000 yards. The input video to the MTI comes directly from the detector on the i-f amplifier strip. The video output of the MTI goes to the switcher mixer.

i. Switcher mixer channel. The switcher mixer channel has two video inputs, one from the MTI and one directly from the i-f strip. The purpose of the switcher mixer channel is to select between the video output of the i-f amplifier and the

output of the MTI. The switcher mixer channel is basically an electronic switch with a variable switchover point. If desired during the initial part of each sweep, the switcher mixer will display MTI video to a range adjustable between 1,000 and 35,000 yards on the PPI's. At that time the switcher mixer switches over and displays the video output of the i-f amplifier on the PPI's. This action is controlled by the MTI ON switch and RANGE control on the acquisition receiver control. If MTI operation is not desired, the switcher mixer channel is turned to OFF, and the switcher mixer channel will display only the video output of the i-f amplifier strips.

j. Video and mark channel. The video output of the switcher mixer channel is amplified and mixed with the necessary marks and gates which make up the PPI display in the video and mark channel. The actual generation and mixing of these marks take place in the mixer channel of the target designator control system, and are discussed in detail in chapter 7. The video and mark channel receives the composite mark signals and the composite video signals and mixes them. The output of the video and mark channel is applied to the four acquisition indicators. The positive video and marks are then amplified and inverted in each indicator and applied directly to the cathodes of the cathode ray tubes. These tubes are intensity modulated so that received signals and marks are displayed as intensified spots on the sweeps.

Section II. RECEIVER FUNCTIONS

43. LOCAL OSCILLATOR (TM 9-5000-26, pages 196 and 197)

a. General. The purpose of the local oscillator is to generate a 3,160-3,560-mc signal which will heterodyne with the target echo to produce the intermediate frequency. The local oscillator tube is a reflex klystron. A special r-f oscillator tube is necessary because of the high frequency operation required. Ordinary vacuum tube oscillators do not oscillate at ultrahigh frequencies because the transit time of the electrons in passing from cathode to plate becomes appreciable in comparison with the period of one oscillation. Crowding together of the tube elements in an effort to reduce the transit time results in increased inter-electrode capacitance which also limits the operating frequency. The klystron overcomes these difficulties by making use of transit time. The local oscillator is a 6BL6 reflex klystron (V5), with an external, coaxially tuned cavity similar to the preselector cavity (see figure 21 for a sectional view of the local oscillator). The oscillator generates a stable r-f c-w signal which, when combined with the received target signal in the signal mixer, produces the 60-mc, i-f signal. A portion of the oscillator output is also coupled into the AFC mixer for the purpose of developing the AFC i-f signal.

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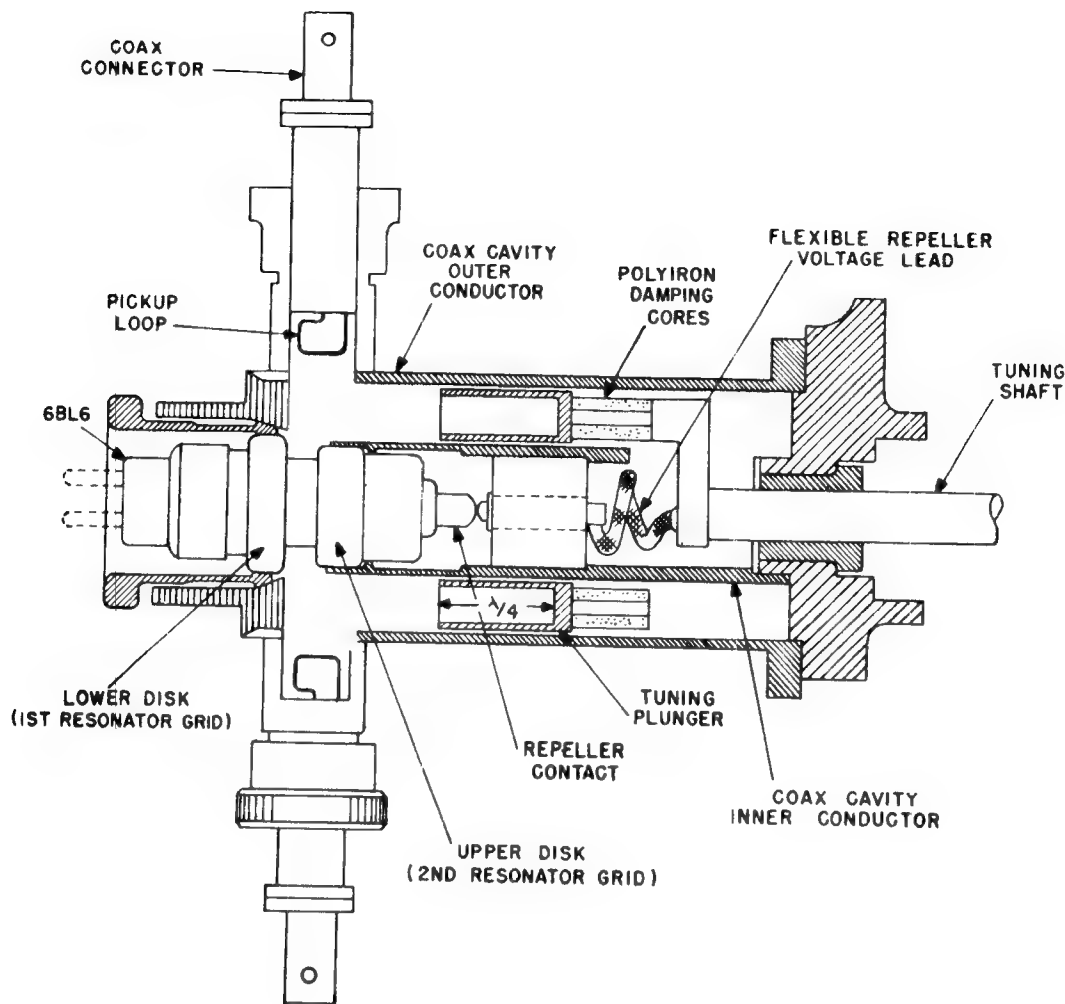


Figure 21. Local oscillator.

b. Operation. The oscillator klystron is mounted within the cavity in such a way that the outer conductor makes contact with the first resonator grid disk. The passage of electrons through the resonator grids tends to initiate oscillation in the coaxial cavity at the natural resonant frequency of the cavity. An oscillating electric field of the same frequency is set up between the grids. Electrons passing through the grids at the time the electric field is zero will undergo no change in velocity, while electrons passing through at other times will be accelerated or decelerated in proportion to the direction and intensity of the electric field. As a result, the electron stream tends to break up into aggregates or bunches in the space between the repeller and the second resonator grid. By adjusting the repeller voltage to a suitable negative value with respect to the cathode, the electron bunches are caused to return through the resonator grid in such phase that the oscillations in the external cavity are reinforced. Power is

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drawn off for use in the signal and AFC mixers through two opposed pickup loops inserted in two cylindrical sleeves perpendicular to the cavity axis. The loops are always oriented by means of keyed connectors to provide uniform coupling throughout the oscillator frequency range. Tuning of the oscillator is accomplished through simultaneous control of two variables, the cavity size and the repeller voltage. The repeller voltage is made more or less negative with respect to the cathode by potentiometer R32. Making the repeller more negative decreases the time required for electron bunches to return to the resonator grids, hence increases the frequency. If at the same time the cavity is made smaller by changing the position of the noncontact plunger, stable oscillation at a new frequency occurs. The tuning plunger and the potentiometer arm are driven by the same AFC motor that tunes the preselector cavity. The whole system is aligned to obtain optimum received signal response when the transmitter magnetron frequency is changed. A powdered-iron ring is located at the closed end of the tuning cup to damp out spurious oscillations which might be initiated in the rear part of the cavity.

c. Tuning. The frequency of the oscillations is primarily dependent upon the dimensions of the cavity, which acts as a resonant circuit. Sustained oscillations are obtained by causing the bunched electrons to be returned to the cavity grids at the proper time. This is accomplished by adjustment of the negative voltage present on the repeller plate. In order to obtain sustained oscillations at a different frequency, it is necessary to adjust both the size of the resonant cavity and the amplitude of the repeller plate voltage. As the volume of the cavity is approximately proportional to the inductance of the resonant circuit, reducing the size of the cavity will reduce the inductance and increase the resonant frequency. At a higher operating frequency it is necessary that the electron bunches be returned to the cavity grids more quickly. Hence, a more negative repeller plate voltage and a smaller resonant cavity are required in order to obtain sustained oscillations at a higher frequency.

d. Adjustment. The two pickup loops which extend into the local oscillator cavity are adjusted to produce 1 ma of crystal current. The front probe controls the signal mixer crystal current, and the rear probe controls the AFC crystal current. These two adjustments interact and, therefore must be adjusted simultaneously in conjunction with the spread and level adjustments.

44. SIGNAL MIXER (TM 9-5000-26, pages 196, 197)

a. General. The signal mixer is a nonlinear device in which the received echo signals are mixed with the output of the local oscillator. An intermediate frequency which can be readily amplified in the i-f stages of the receiver is produced. When two frequencies are mixed in a nonlinear device such as the signal mixer, the resultant includes many frequencies. Among these are two original frequencies,

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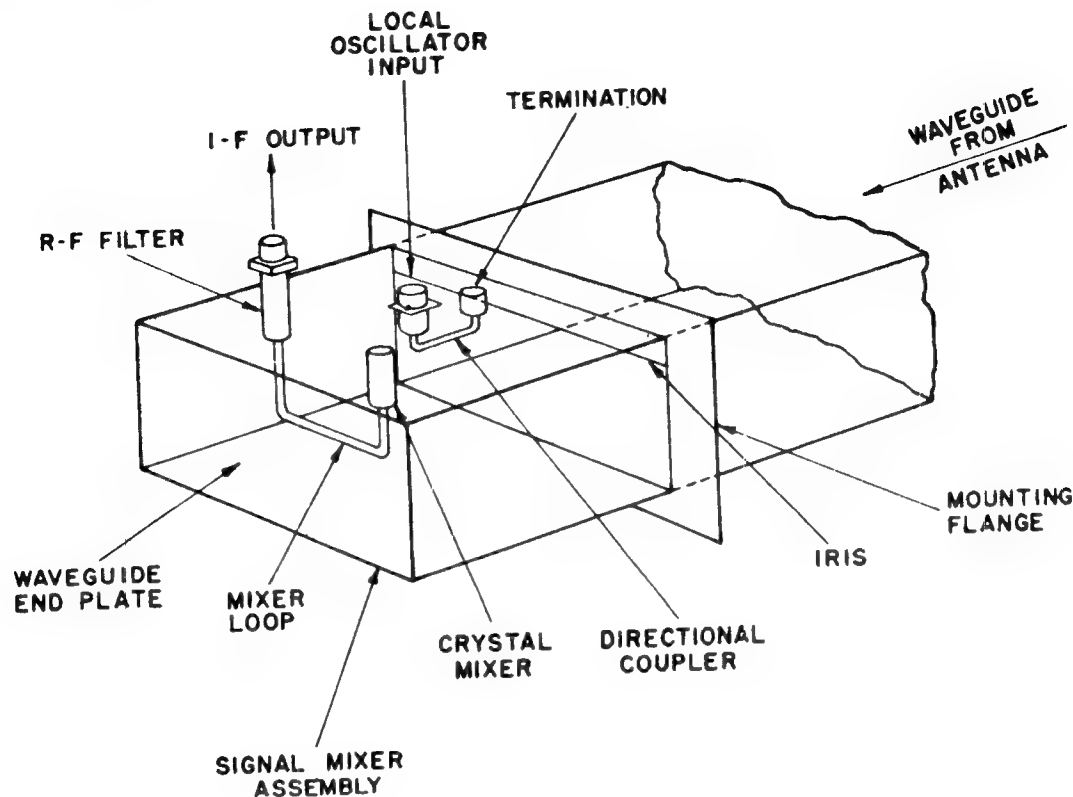


Figure 22. Signal mixer.

the sum of the two frequencies, the difference between them, and sum and difference frequencies of the various harmonics. The higher frequencies serve no useful purpose after mixing and are removed by filtering in the output circuit, which is designed to pass only the 60-megacycle intermediate frequency. The signal mixer is a closed-end section of waveguide (fig 22). It contains the elements necessary for mixing the local oscillator output with the received signal and extracting the 60-megacycle difference frequency. Local oscillator energy is introduced into the mixer through a coaxial cable which feeds a directional coupler inside the waveguide. The coupling element is a U-shaped continuation of the center conductor and extends about one-quarter wavelength through the waveguide in a direction parallel to its main axis. It is terminated in a powdered-iron core, making the coupling device unidirectional. Most of the energy not coupled into the mixer is absorbed by the cylindrical powdered-iron core which terminates the coupler. The coupler is made directional toward the mixer to prevent local oscillator energy from leaking out to the antenna system. However, the maximum directivity is limited by the fact that approximately 10 percent of the local oscillator energy is reflected from the mixer loop. The mixer assembly is coaxial in

construction, with the center conductor forming a transverse pickup loop within the waveguide. This loop is terminated at one end by the crystal mixer. Impedance matching of the waveguide to the crystal is accomplished by the dielectric case of the crystal holder. The i-f connector is equipped with a concentric powdered-iron core. This core acts as an r-f choke to eliminate all but the 60-megacycle, i-f signals from the output. Conversion loss through the signal mixer is about 8 decibels. The local oscillator develops more than 100 milliwatts of power although only 1 milliwatt is required for optimum mixing. Ten milliwatts are drawn from the local oscillator, but most of this supply is dissipated in the termination for the local oscillator input.

b. Operation. The crystal mixer loop acts as a receiving antenna and picks up the r-f energy in the crystal mixer cavity. This energy consists of the mixed output of the local oscillator and the received signal. The crystal rectifies this mixed signal, allowing the positive portion and a small part of the negative portion of the voltage waves to cause current to flow in the output circuit. The received signal mixes the local oscillator waveform at this stage. This mixing causes a new signal at a frequency of 60 megacycles (local oscillator frequency minus the received signal frequency) to appear. The output circuits of the crystal mixer offer a high impedance to the r-f frequencies contained in the local-oscillator and received signals, while a low impedance path for these frequencies exists in the capacitance of the r-f filter. Thus, the r-f components of the mixed signals in the mixer loop are shunted to ground in the r-f filter, leaving only the 60-megacycle intermediate frequency to affect the next stage.

c. Crystal construction. The crystal used in this mixer is in the IN 28. It consists of a small crystal of almost pure silicon with a brass plate on one end and a fine tungsten wire (cat's whisker) pressing lightly on the other. It is at this junction of tungsten to crystal that rectification takes place. Since the transit time of the electron stream through a tube is an appreciable fraction of the time of one cycle of r-f energy, a diode would not perform satisfactorily in this application. The receiver's sensitivity is dependent in large measure on the crystal's performance. The crystal must be extremely sensitive to the small amount of energy existing in the i-f signal and must not generate appreciable noise at frequencies near the i-f frequency.

d. Care of crystals. Since extreme sensitivity is required of the crystal, it is fragile both electrically and mechanically. The crystal should not be subjected to mechanical shock by dropping or to excessive current. Shock may cause the tension with which the cat's whisker presses against the silicon crystal or may cause the area of the junction to change. Excessive current causes heat at the tungsten to silicon junction which may burn the thin tungsten wire or melt the silicon crystal, changing its crystalline structure. Rules to follow are listed below.

- (1) Do not drop the crystal.
- (2) Keep the crystal wrapped in its lead foil case until ready to use. This foil protects the crystal against the strong r-f fields usually present around radar equipment.
- (3) Ground yourself to the metal framework of the set before unwrapping or changing a crystal. The static charge built up in the human body is of high enough potential to seriously damage the crystal if allowed to discharge through it.
- (4) Do not exceed the recommended crystal current when adjusting the local oscillator to crystal mixer coupling. The crystal current as read in the metering circuit is made up of the current flowing as the result of all the frequencies present. These are the local oscillator frequency, the received signal frequency, the sum of these two, and the intermediate frequency. However, the received signal normally contains very little energy and is present for only a short time during each cycle. Therefore, the crystal current is almost entirely due to the output of the local oscillator.

45. CRYSTAL USE AND CARE

a. Checking of crystals. If in the adjustment of the local oscillator coupling it is not possible to obtain the rated crystal current, the trouble may be that the local oscillator is not oscillating strongly enough or that the rectifying properties of the crystal have deteriorated. The crystal should then be checked with a crystal checker. The crystal should exhibit a front-to-back resistance ratio in the neighborhood of 29 to 1. If a crystal checker is not available, the front-to-back resistance ratio can be checked using the ohmmeter of the TS-352U. However, when the TS-352U is used, a scale lower than RX100 should not be used. A lower scale will damage the crystal because of excessive meter current. If the front-to-back resistance rating of the crystal is satisfactory, the crystal can be reinstalled, and most likely the local oscillator circuit is at fault and should be checked. If it is found that the crystal has to be replaced frequently, even though rules referred to above have been strictly adhered to, the cause may be a faulty TR tube. The TR tube should then be tested and, if found faulty, replaced. The TR tube can cause the crystal to become defective because, the transmitter, when firing, sends a tremendous burst of energy down the waveguide. The TR tube is designed to keep all but a minute portion of this energy from reaching the crystal mixer. However, as the TR tube ages, it becomes increasingly more difficult for the transmitter energy to ionize the TR tube, and a considerable amount of transmitter energy passes through the TR tube to the crystal mixer before the voltage in the TR tube builds up sufficiently to ionize the TR tube. Even though this energy is present for only a fraction of the pulse duration, the energy can be of sufficient amplitude to damage the crystal if the TR tube has deteriorated to any degree.

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b. Selection of crystals. As previously stated, the receiver is designed to have a high signal-to-noise ratio. One of the main sources of noise, and consequently one of the elements which causes a lowering of the signal-to-noise ratio, is the crystal mixer. The front-to-back resistance ratio is only one measure of the crystal's quality. The other is the noise-free operation of the crystal. All crystals generate noise in operation, but some generate less noise than others. The repairman must select a crystal that is relatively noise-free, although no test equipment other than the receiver itself, is available to check the crystal. The repairman must try several crystals in the crystal mixer, carefully noting the sensitivity of the receiver with each crystal and then placing that crystal in the crystal mixer which gives the best receiver sensitivity. Since mechanical shock and excessive current also change the crystal's noise characteristics, the rules cited in 44d apply here.

c. Resistive card. A spring-backed plunger protrudes from the right side of the signal mixer cavity and is connected internally to a phenolic card covered with powdered carbon. This diamond-shaped card normally rests against the side of the waveguide and is therefore inactive. Pushing the plunger places the card near the center of the signal mixer waveguide, with the resultant absorption of about 7 or 8 decibels of energy. In this manner, the preselector cavity of the receiver will have a more resistive impedance to work into, permitting a better adjustment of this cavity. It is necessary, then, to push the plunger when the preselector is being tuned so that it will be accurately adjusted over its entire frequency band.

46. CURRENT METER (TM 9-5000-26 pages 196, 197)

Meter M2 on the r-f coupler meter panel has been incorporated to measure the amount of signal mixer and AFC mixer crystal current as well as limiter current in the AFC. When the meter is switched (S9 position 3) to measure signal mixer crystal current, current flows from crystal CR2 to ground, from the grounded side of the meter M2, through the meter, through the 60-mc filter consisting of L1, L2, L3 and C1, C2, C3, through the signal mixing loop, and back to the other side of the crystal. The purpose of C4 across the meter is to bypass any r-f energy that might not be filtered out by the previous filter. Resistor R24 is a meter shunt and also serves to complete the circuit when the meter is switched to another circuit.

47. I-F PREAMPLIFIER CHANNEL

a. General (TM 9-5000-26, page 207). The purpose of the i-f preamplifier channel is to build up the strength of the received echoes before they are sent over the long cable to the fire control van. Once inside the trailer, this 60-mc, i-f signal is further amplified in the i-f amplifier channel and then detected. The

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resulting video information is further amplified in the i-f amplifier channel and then detected. The resulting video information is further amplified and finally applied to the acquisition indicators.

b. Block diagram discussion. The acquisition i-f amplifier channel is located at a considerable distance from the acquisition antenna. Coaxial cable transmission of the output of the signal mixer would result in prohibitive attenuation and mixing of a relatively large amount of noise with the intermediate frequency. Therefore, the amplitude of the i-f signals must be increased before the signals are conveyed to the receiver circuits in the fire control van. This amplification is accomplished by the i-f preamplifier channel. Since maximum power gain through the preamplifier channel is about 1,000,000 times and since the input signal strength is extremely low, special design precautions have been taken to hold random circuit and tube noises to a minimum. Triodes are known to be less noisy than pentodes. Therefore, the first two tubes are triodes operated in grounded-grid circuits for frequency stability. Heavy loading eliminates the triode's natural tendency to oscillate. The gain through each stage is not high, but the use of the first two stages raises the signal to a satisfactory level for further amplification with low noise content. The remaining three stages are high-gain pentodes connected as conventional i-f amplifiers. The coupling transformers between stages are all tuned to 60 megacycles per second, plus or minus 0.5 megacycle. A bandwidth of 1.75 megacycles per second at the 3-db points is obtained by heavily loading the transformer windings. The gain of the i-f amplifier channel is controlled by a composite voltage applied to the control grids of tubes V3, V4, and V5. This voltage is obtained from the sensitivity time control channel (STC) and is controlled at the acquisition receiver control. Part of the control voltage is a negative d-c potential, variable between 0 and 20 volts by means of GAIN control. This voltage permits adjustment of receiver gain over the entire sweep range. Superimposed on this d-c voltage is the negative STC signal, which occurs during the first part of the sweep. The amplitude, shape, and duration of this signal are all adjustable. The function of the STC signal is to decrease the gain of the i-f preamplifier channel for echoes received from nearby objects and to permit gradual restoration of the operating level for echoes occurring farther out in range on the radial PPI sweep. The effect is to insure that the presentation on the PPI's will have a uniform brilliance over the full range. The over-all gain is set for clearest display of targets at extreme range without causing nearby targets to become fuzzy due to excessive signal amplitude.

c. Detailed operation of V1 and V2 (TM 9-5000-26 page 210). C4 is a coupling capacitor for the 60-megacycle signal which is applied to the cathode of V1. This stage employs a grounded-grid triode in order to minimize induced grid noise and interception noise. The input circuit to V1 is tuned to 60 megacycles by Z1. Z1 provides the proper input impedance. Resistor R1 provides the correct bias for V1. C5 and C6 are r-f bypass capacitors. Current flows through

V1 from ground through R1 and Z1, through the tube to the primary of T1, and through decoupling resistors R2 and R16. From R16, current flows through the plate decoupling network consisting of C22, C21, R18, R19, and R20 to the 150-volt power supply, to ground. T1 is overcoupled, reflecting a heavy load to the plate of V1 and thereby broadening the bandpass. The signal is coupled by T1 to V2. V2 is also a grounded-grid triode amplifier, assisting V1 in attaining a high signal-to-noise ratio. R3 produces the necessary bias for V2, and L4 develops the signal applied to the cathode of V2. C8 is an r-f bypass capacitor and C7 is a blocking capacitor. R4 detunes L5 to widen the bandpass of V2. The signal passes through C9, which is a coupling capacitor, to the grid of V3. The input to V3 is sharply tuned to 60 megacycles by means of Z2. Z2 is designed to resonate with grid and distributed capacitance at 60 megacycles.

d. Final three stages. The remaining stages are high-gain pentodes connected as conventional i-f amplifiers. The coupling transformers, T2, T3, and T4 are all tuned to 60 megacycles per second center frequency. A bandwidth of 1.75 megacycles at the 3-db points is obtained by heavily loading the transformer windings. T2 and T3 are loaded by resistors R7, R8, R12, and R13, respectively. T4 is shunted by resistor R21 and is loaded by the low impedance of the coaxial line to the i-f attenuator. V3, V4, and V5 each have a 220-ohm resistor in the cathode circuit for proper cathode bias and an r-f bypass capacitor for stabilization of the cathodes. In the output circuit of T4, C20 is factory-adjusted to match the impedance of the coaxial cable going from the i-f preamplifier to the i-f attenuator. The gain of the preamplifier, and hence of the receiver, is controlled by a composite voltage applied to the control grids of the final three stages, V3, V4, and V5. This voltage comes from the sensitivity time control channel (STC) and is controlled at the acquisition receiver control.

48. ATTENUATOR (TM 9-5000-26, page 215)

a. General. The purpose of the attenuator is to compensate for wide variations in gain of the 12 stages of the i-f preamplifier channel and the i-f amplifier channel due to aging of tubes and changes in value of circuit components. The signal from the i-f preamplifier channel travels through the 250-foot cable to the van and experiences a loss of about 9 decibels. This 60-mc, i-f signal is then applied to the i-f attenuator located on the carrier generator chassis. The i-f attenuator allows the insertion of 20-40 decibels of loss in 5-db steps. The attenuator contains several resistance networks connected to a switch. These pi-type networks introduce 5-, 10-, 15-, or 20-db loss. In addition, there is fixed attenuator consisting of R32, R33, and R34, which adds an additional 20-db loss. As a result, the total loss through the attenuator is always at least 20 decibels but may be increased to 40 decibels in 5-db steps. The output of the attenuator is connected to the i-f amplifier channel.

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b. Adjustment. The attenuator is adjusted with the test amplifier connected to the acquisition video test point on the video and mark mixer chassis and with the receiver gain turned clockwise to the maximum. With these connections made and the transmitter on, adjust the attenuator until the receiver noise is $1/2$ the height of the main pulse.

49. I-F AMPLIFIER CHANNEL (TM 9-5000-26, page 211)

a. General. The purpose of the i-f amplifier channel is to amplify the 60-mc i-f signal to a usable level and then detect the modulation envelope. The first six stages are all broadband i-f amplifiers; the seventh stage is a power amplifier, and the eighth is a dual detector. The bandwidth of the i-f amplifier channel is 4 megacycles. The two outputs from this dual detector are negative bypass video and positive for the MTI circuits video. The bypass video is sent through the switcher mixer channel and later presented on the acquisition indicators, while for the MTI circuits, video is sent to the MTI circuits and finally to the indicators if needed.

b. Amplifiers V1 through V6. The operation and function of the first six stages are so nearly alike that they will be discussed collectively. Identical transformers are used to provide inductive interstage coupling. Both plate and grid circuits are tuned to 60 megacycles, plus or minus 1.5 megacycles. The primary of the input transformer T1 is series-tuned by capacitor C1 in order to match the low-line impedance and still maintain a step-up relationship through the transformer. Both the primary and secondary windings of the interstage transformer are tuned to 60 megacycles by interelectrode and distributed capacitances. Resistive loading of these windings broadbands the transformers, increasing the fidelity of signal reproduction. Control grid bias of +35 volts is obtained from the voltage divider, R59 and R60, connected between +150 volts and ground. Capacitors C2, C4, and similar capacitors in other stages place the lower end of the transformer windings at signal ground. Resistors and capacitors between pin 4 of plug P2 and the screen grids and plates serve to decouple all the plate and screen grid leads from each other and from the voltage source. The capacitors and resistors in the control grid circuits between the voltage divider and the grids perform a similar function. Each stage is designed to prevent blocking by signals which have a duration of more than one microsecond. This feature, called back bias, minimizes PPI ground clutter and aids MTI operation. The effectiveness of c-w jamming is also reduced by this feature. Back bias is obtained through use of a large resistor and a small bypass capacitor in each cathode circuit. Because of the relatively high voltage developed across the 12,000-ohm cathode resistors, the tubes operate near cutoff in the quiescent state. The values of the cathode resistors and associated bypass capacitors are such as to hold the cathode voltage constant for signals not greater than one microsecond in length. For signals up to one microsecond, the capacitor

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bypasses the resistor, and the bias of the stage is unchanged. However, only the initial 1-microsecond portion of a longer signal is bypassed. For the remainder of the signal, degeneration is introduced by the effectively unbypassed resistor, and the gain of the stage is reduced. This system is known as a back-bias method of gain control since the controlling bias is determined by the amplitude and duration of the applied signals. Maximum gain cannot be obtained in these stages because of this type of operation.

c. Power amplifier V7. This stage is a conventional high-gain power amplifier. It is inductively coupled to the preceding stage and to the detector which follows it. Both plate and grid circuits are tuned to the intermediate frequency by the interelectrode capacitance and the distributed capacitance of the transformer windings. A load resistor is placed across these windings for broad-banding. Normal cathode bias is obtained by means of R49 and C39. As in the first six stages, the plate and the screen grid are decoupled from the voltage supply. This is done by R53 and C38.

d. Dual detector V8. This stage is a double-diode detector operated to obtain two video outputs of opposite polarity. Transformer T8 is connected to the plate of one section and to the cathode of the other section. Each section conducts when the applied signal renders the plate more positive than the cathode. V8B, the bypass-video detector, has its plate-load resistor physically located in the switcher mixer. This is a 25,000-ohm variable resistor, R13, labeled BYPASS GAIN. A similar resistor, MOD LEVEL, in the carrier generator, is the load for the MTI video detector, V8A. The L-C filters in the load circuits serve to eliminate any 60-megacycle component remaining in the detected video signal. The video used in the MTI requires more complete filtering than does bypass video. For this reason, a pi-section filter consisting of L2, C47, and C48 is employed. The capacitors in both plate circuits charge and discharge in an attempt to maintain a constant current through the load resistors. The inductors, L1 and L2, oppose changes in current and so help to eliminate ripple. Their value must be small enough to prevent broadening of the video pulses.

50. PRESELECTOR (TM 9-5000-26, pages 196, 197)

a. General. The purpose of the preselector is to allow only target echoes at the magnetron frequency to enter the receiver. This cavity is mounted as an extension of the waveguide between the TR tube and the signal mixer. It is a resonant cavity, cylindrical in shape, which is tunable over the frequency of the magnetron.

b. Operation. In its operation, it can be likened to a parallel resonant circuit offering maximum impedance to the resonant frequency. It is tuned by the receiver tuner motor B3 to the frequency of the magnetron, and hence provides

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maximum transfer of signal energy at the magnetron frequency. Signals at frequencies other than the transmitter frequency are attenuated. The design of the preselector is such that maximum attenuation is experienced by the image frequency (local oscillator frequency plus 60 megacycles). This action provides wanted-to-unwanted signal discrimination and makes electronic countermeasures less effective against the acquisition radar. The preselector also reduces interference from other sets operating on adjacent frequencies.

c. Circuit analysis. The tuned cavity acts as a shunt-coupling resonant tank to the wanted signal frequency. Tuning is affected by double-cup, noncontact plungers moving in the space between the inner and outer cylinders. Spurious oscillations in the portion of the cavity below the tuning cups are prevented by the use of a powdered-iron damping ring. To insure that the cavity resonates only in the desired mode, the outer conductor is pierced at the waveguide end by two vertical slots 90° apart. The preselector is coupled to the waveguide through these two windows. Preselector insertion loss is in the order of 0.5 decibel for signals at the resonant frequency and as high as 45 decibels for the image frequency.

d. Field adjustment. The preselector cavity is tuned through a gear system driven with the local oscillator cavity and the local oscillator repeller plate potentiometer. The preselector must be adjusted to track with the local oscillator each time the local oscillator tube is replaced. This alinement is done by disconnecting the gearing between the oscillator and preselector cavity and then adjusting the preselector for maximum signal on the PPI's. This gearing can be released by loosening a nut just above the gear that drives the preselector cavity. This will cause the gears to separate, and the preselector alone may be tuned manually. A very effective means of viewing the results is to connect an oscilloscope to the test synchronizer and test video jacks on the r-f coupler, and then connect the acquisition video signal into this test video cable inside the acquisition cabinet behind and below the power panel. To make this connection, disconnect the coaxial cable on J-19 and connect the cable above (P20) to J19. This way the output of the receiver may be viewed on the oscilloscope by the person making the preselector adjustment. This procedure has the advantage, also, that an A-scope presentation is used, and the persistence of the scope will not cause errors in tuning. When the preselector is tuned, a spring-loaded plunger on the signal mixer should be depressed to improve the tuning accuracy (par 45c). Due to the very narrow bandpass of the preselector, its tuning is very critical, and even slight errors will prevent long-range targets from being seen.

51. LOCAL OSCILLATOR POWER SUPPLY (TM 9-5000-26, page 209)

a. Power supply outputs. The local oscillator power supply furnishes the following voltages:

- (1) -800 volts as keep-alive voltage to the TR tube, V4.

- (2) -625 volts to the circuit which controls the local oscillator repeller plate voltage.
- (3) -325 volts to the cathode circuit of the local oscillator.

b. Rectifier and regulator operation. A dual-diode tube, V1, operates as full-wave rectifier. When the ACQUISITION POWER switch is operated to the ON position, 120 volts alternating current, phase A to neutral, is applied to the rectifier circuit. Transformer T1 steps up this voltage to 1,600 volts, which is applied to the plates of V1. Since the cathode of V1 is grounded, the -800 volts direct current is taken from the center tap of T1. This output is then filtered to remove the 800-cycle ripple frequency. Filtering is provided by an L-C circuit consisting of L1, C1, C2, and C3. A filament voltage of 6.3 volts is available at the secondary of transformer T2, but resistors R1 and R2 reduce the a-c voltage applied across the filaments of V1 to 5 volts. The filtered d-c voltage of -800 volts is developed across a combination voltage regulator and voltage divider network consisting of regulator tubes V2 through V6 and resistors R3 through R8. V2 and V3 are 150-volt regulator tubes. V4, V5, and V6 are 108-volt regulator tubes. V7, a 75-volt regulator tube, insures that the repeller plate of the local oscillator will always be at least 75 volts more negative than the cathode.

c. M1. Meter M1 is used primarily to check power supply voltages. This meter is located on the left side of the r-f coupler meter panel. This circuitry is shown on pages 196 and 197 of TM 9-5000-26. Eight ranges are covered by the selector switch S8 just below the meter. Position 1 measures average magnetron current, while all other positions measure supply voltages. Position 6 is of interest because the +320 voltage on this position appears after the acquisition 15-minute delay has elapsed. Positions 7 and 8 give a check on the local oscillator power supply. The various resistors associated with switch S8 are meter multipliers to allow the meter to measure a wide range of voltages.

d. Spread and level circuits. The spread and level controls are incorporated to adjust the repeller plate voltage on the local oscillator klystron. Two problems are overcome here. First, the plate voltage is very critical with respect to frequency, so R32, the repeller plate potentiometer, taps off higher negative voltage as the frequency is increased. Secondly, the voltage level and spread for the frequency band needed varies with each klystron used and must be adjusted when the tube is changed. No adjustment of the voltage applied across repeller plate potentiometer R32 is possible by means of the spread (R31) and level (R30) controls. These controls are adjusted so that the repeller voltage will be proper for operation throughout the frequency range of the local oscillator, and -625 volts is applied to terminal 1 of R30A, and -400 volts is applied to terminal 1 of R30B. Movement of the brush arm of R30 (the level control) toward terminal 1

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increases the resistance introduced in the circuit by R30A and decreases the resistance of R30B. A greater voltage drop will appear across R30A, and a smaller voltage drop will appear across R30B. As a result, terminal 3 of R30A and terminal 3 of R30B will become less negative. The potential difference between these two points is maintained at 108 volts by voltage regulator tube V10. Adjustment of level control R30 determines at what level (between -625 and -400 volts) this 108-volt potential will be. The entire 108 volts is not applied across the 40,000-ohm repeller plate voltage potentiometer R32, as limiting resistors R28 and R29 and spread control R31 are in series with that potentiometer. Movement of the brush arms of R31A and R31B will vary the total resistance in series with R32 and, hence, will vary the amplitude of the potential applied across R32. This potential difference may be varied within limits of 41 volts and 68 volts. Test points TP1 and TP2 are provided for monitoring the voltage applied to repeller plate voltage potentiometer R32. Test point TP4 is provided to monitor the voltage on the local oscillator repeller plate. The voltage at this point can be varied from -420 to -572 volts, depending on the setting of the spread and level controls. Note that these test points are on the r-f coupler meter panel, while those test points associated with the regulator tubes V2-V6 are on the local oscillator power supply chassis.

e. Spread and level adjustment.

- (1) Adjust the receiver tuner to the middle of its range.
- (2) Adjust the level control to obtain maximum signal mixer crystal current as read on meter M2 in position 3.
- (3) Tune the receiver tuner to its extreme high frequency end (plunger all the way in).
- (4) Turn the receiver throughout the entire band and see that the crystal current is fairly uniform. The value of current should be around 1 milliamperere. However, 0.5 to 1.5 milliamperere is acceptable. The maximum current is determined by the local oscillator coupling loops.

Section III. RECEIVER TUNING CIRCUITS

52. AUTOMATIC FREQUENCY CONTROL (TM 9-5000-26, page 214)

In order to insure reception in complex radar systems, the local oscillator must operate with a constant frequency difference between it and the magnetron, the received signal having the same frequency as the transmitted signal. This difference is equal to the intermediate frequency. Since drift characteristics between magnetrons and local oscillators frequently vary, some automatic means

of maintaining a constant difference is necessary. The system of circuits which accomplishes this objective is known as the automatic frequency control (AFC) channel. In the Nike I system there is an additional reason for incorporating AFC. The acquisition transmitter magnetron is tunable throughout the range of 3,100 to 3,500 megacycles. When the magnetron frequency is changed, the AFC system acts to restore the 60-mc frequency difference between the local oscillator and the magnetron. Further, since there are two local oscillator frequencies, one above and one below, which will produce a 60-mc beat frequency, the system is arranged to tune the local oscillator only to the higher frequency. An i-f signal is developed in the AFC mixer from the combination of a small portion of the magnetron energy and the local oscillator output. The i-f pulses are then fed to i-f amplifiers of the AFC channel, where any deviation from 60 megacycles is converted through a sequence of operations to a 400-cycle, a-c voltage whose phase and amplitude are determined by the direction and amount of the initial frequency deviation. This a-c control voltage is conveyed to the low-power servoamplifier. Here, its power level is increased and the resulting signal is applied to the AFC tuning motor in the receiver tuner. The tuning motor adjusts the size of the preselector and local-oscillator cavities and the magnitude of the local-oscillator klystron V5 repeller voltage. As the frequency deviation from 60 megacycles is diminished, the correction signal around the control loop is also diminished until the local oscillator is once again set 60 megacycles higher than the magnetron frequency. The complete AFC channel can therefore be called a closed-loop servo system.

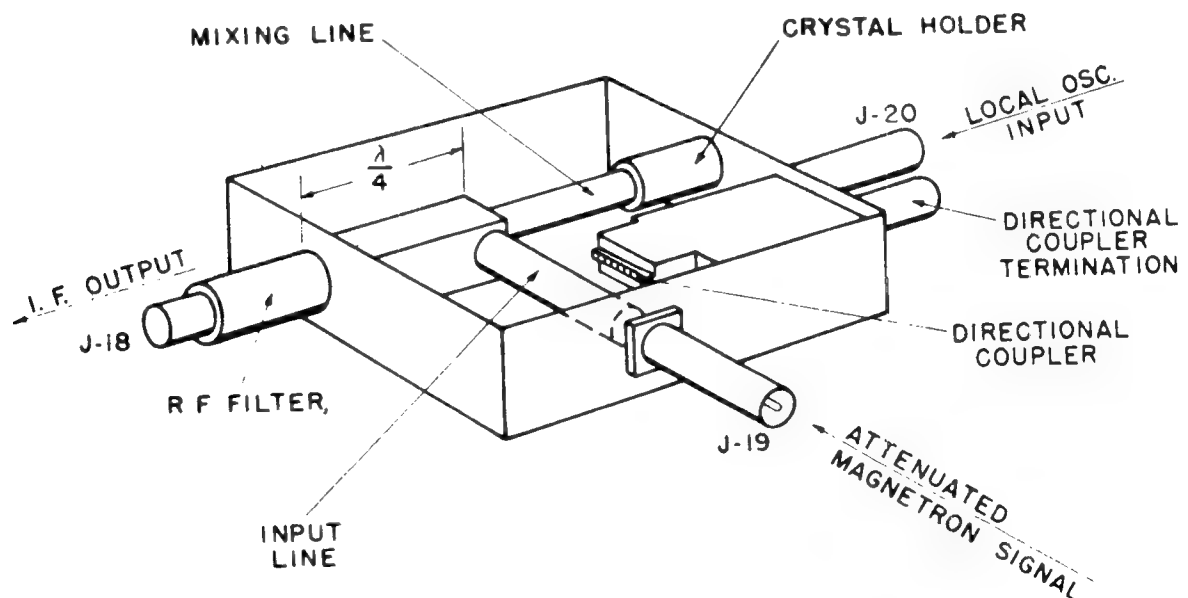


Figure 23. AFC mixer.

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53. AFC MIXER (TM 9-5000-26, pages 196, 197)

a. General. The AFC mixer is similar in operation to the signal mixer. It is located on the left side of the receiver tuner assembly. The AFC mixer (fig 23) mixes a small portion of the transmitted signal with the local oscillator output and detects the resulting i-f frequency. The detected signal is applied to the i-f amplifiers of AFC channels. Note that only transmitter output and not received signals are mixed with the local oscillator output in the AFC mixer. Thus, signals from other transmitters cannot enter the AFC mixer to give a false i-f signal.

b. Operation. The small portion of transmitted energy used in the AFC mixer is obtained through a coupling window located in the narrow side of the waveguide between the magnetron and the ATR tubes. This window couples in a metal cylinder. A pickup loop within the cylinder transfers the energy to a coaxial cable, which sends it to the mixer. Energy from the local oscillator is also brought into the mixer by coaxial cable. The crystal is mounted in one wall of the mixer at the end of the mixing line. A dielectric sleeve in the crystal holder provides capacitance for resonating with the crystal inductance in order to obtain optimum matching. A quarter-wave, low-impedance line is used to shunt the 50-ohm, transmitter, r-f input line where it couples to the mixing line. This gives the mixer a broad bandpass. The shunt impedance is a square post which projects into the mixer box. This post also supports the powdered-iron core, which acts as an r-f filter in the i-f output line. The directional coupler couples energy from the local oscillator into the mixer line. A powdered-iron core at the termination of this coupler is used to attenuate the local oscillator energy to the desired level of one milliwatt. Matching is obtained through the use of dielectric rings which also serve to center the inner conductor of the local oscillator input line. When two frequencies are mixed, the result is the development of many frequencies. The two original frequencies, the sum of the two input frequencies, the difference between the two input frequencies, and the sum and difference of the various harmonics are present. The input frequencies, the sum frequency, and the harmonics serve no purpose and are filtered out by the r-f filter at the i-f output line. The i-f signals are detected and sent through a coaxial cable to the i-f amplifiers of the AFC channel.

c. AFC crystal current meter M2. The meter M2 which measures signal mixer crystal current, AFC crystal current, and AFC limiter current is located on the meter panel in the r-f coupler. When the meter is switched to AFC crystal current (S9 to position 2), current flows from the grounded side of CR1 through this crystal, through transformer T1, through filter R15 and C10A, to the positive side of meter M2 from the negative side of the meter, to ground, back to CR1. Capacitor C4 bypasses any intermediate frequency around the meter.

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54. I-F AMPLIFIERS (TM 9-5000-26, page 214)

The i-f signals are brought into the AFC unit at J1. The coupling transformers, T1 through T4, are fixed-tuned to resonate at 60 megacycles with their stray capacitance. Resistors R28 and R5 through R11 load the transformers to provide broadbanding. The R-C network in the cathode circuit of each of the amplifiers develops the proper self bias for class A operation. Capacitors C10 through C12 are plate and screen bypass capacitors which provide a ground for the i-f signals and prevent interaction between the stages. Capacitor C1 provides an a-c ground for the primary of transformer T1 and prevents the intermediate current from reaching the AFC crystal current meter. The over-all bandwidth of these amplifiers is 10 megacycles. This is a usable bandpass since the auto search circuits do not take over until this is exceeded. Crystal current flows in the input cable to the AFC. The path of this crystal current is from the AFC mixer crystal to ground; from ground through the meter switch S9, or through R26 if the meter is in OFF, through meter M2, through a filter consisting of R15, C10A, C1, through the primary of T1, through the cable to the AFC mixer back to the crystal CR1. Resistor R25 across M2 is a meter shunt. The output of the i-f amplifiers is 1.3 microsecond pulses of intermediate frequency.

55. LIMITER (TM 9-5000-26, pages 196, 197, and 214)

The limiter stage V4 is used to eliminate the effect of variation in the amplitude of the i-f signals so that the discriminator may respond only to frequency variations. The tube is operated at a lower plate voltage than the preceding stages and saturates easily. Grid leak bias is used (R22 and C4), in addition to cathode bias, causing the tube to operate below cutoff for high-amplitude signals and near or above cutoff for low-amplitude signals. The effect of this type of operation is to limit the amplitude of strong signals while amplifying weaker signals from the i-f strip. Actually, uniform output from the limiter is obtained over a wider frequency range than the i-f strip bandpass, since signals beyond the shoulders of the i-f response curve are strong enough over a short frequency span to saturate the limiter. The negative grid leak bias developed at the limiter grid is also used to control auto search tube V11A. Limiter action may be monitored in two ways. Current in the limiter grid circuit may be read on meter M2, or the voltage developed as a result of limiter current may be measured on TP3. The path of the grid current for this stage is from the grid of V4 through the secondary of T4, through the filter consisting of C4, C37, C10B, and R17, R22. This current, then, flows through meter switch S9, meter M2, to ground. From ground the current flows through R4 and back to the cathode of V4. A limiter current of at least 10 microamperes is needed if the AFC is to operate correctly. This limiter current may be adjusted by varying the inputs to the AFC mixer. The output of limiter V4 is 1.3-microsecond pulses of constant amplitude.

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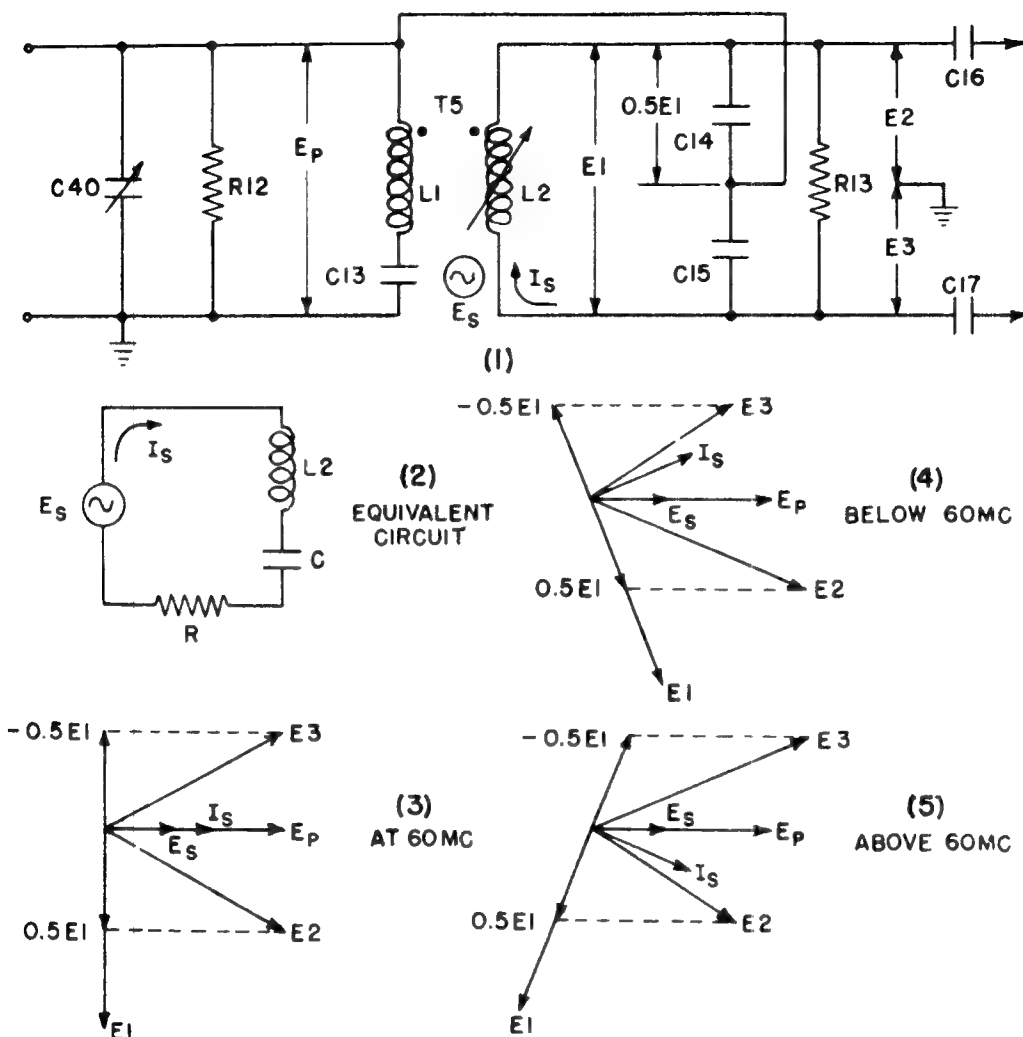


Figure 24. Discriminator.

56. DISCRIMINATOR V5

a. General. The discriminator converts frequency variations of the i-f signal into video pulses 1.3 microseconds in duration. The amplitude of the pulses is proportional to the deviations of the intermediate frequency from 60 megacycles. Figure 24 shows the vector plot of altitude and direction of the pulse. The polarity of the pulses is determined by the direction of the frequency deviation. If the input frequency is less than 60 megacycles, the discriminator output pulses are positive; if it is more than 60 megacycles, the output is negative. The pulses developed by the discriminator as a result of an error in the intermediate frequency are used to produce an a-c voltage which drives the tuning motor. This retunes the local oscillator to correct the error that existed in the intermediate frequency. The limited i-f signals from tube V4 are

coupled to the discriminator detector diodes, V5A and V5B, through transformer T5. The secondary of the transformer is a series-resonant circuit, tuned to 60 megacycles. As long as the input frequency is 60 megacycles, the inductive and capacitive reactances present in the secondary circuit are equal, and voltages that are of the same magnitude appear at opposite ends of an essentially resistive circuit. These voltages are applied to the diodes and cause equal and opposite currents to flow in the detector circuit. When equal currents flow through the diodes, no output pulses are produced by the discriminator. If the input frequency varies from 60 megacycles, the inductive and capacitive reactances in the tuned circuit are not equal. The resultant voltages which then appear at opposite ends of the circuit differ in amplitude and in phase. Unequal currents are caused to flow in each half of tube V5. This results in the development of output pulses. The amplitude of these video pulses is proportional to the magnitude of the deviation of the intermediate frequency from 60 megacycles, and the polarity is determined by the direction of the frequency deviation.

b. Operation.

- (1) The primary and secondary windings of transformer T5 (fig 24) comprise two tuned circuits, both resonant at the intermediate frequency. The input voltage E_p varies little with frequency variations, hence the voltage E_s induced in the secondary is relatively constant. The transformer is wound so that E_s is in phase with E_p . The voltage E_s is small compared with E_p because of loose coupling, but as the secondary of T5 is a series-tuned circuit, a relatively large current I_s flows in the secondary. This current develops voltage E_1 across the resonant secondary circuit. E_1 is large compared with the driving voltage E_s . As the values of C14 and C15 are equal, and as the voltage across the two capacitors is E_1 , then a voltage equal to $0.5 E_1$ will appear across each capacitor.
- (2) The magnitude of the current I_s is determined by the total impedance in series with E_s . At the resonant frequency, this impedance is a pure resistance and is minimum. In addition to the pure resistance, inductive reactance is present above resonance and capacitive reactance below resonance. Thus, both the magnitude and phase of E_1 change when the frequency of E_p changes. The voltages E_2 and E_3 , from the plates of the diodes to ground, are related to E_p and E_1 as follows:

$$E_2 = E_p + 0.5 E_1$$

$$E_3 = E_p - 0.5 E_1$$

The changes in E_2 and E_3 with frequency produce the discriminator action of the circuit.

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- (3) Figure 24(2) is in the equivalent circuit of the secondary of transformer T5. Coil L2 is the secondary winding of T5. Capacitor C represents C14 and C15. Resistor R represents the d-c resistance of the coil as well as the shunt resistor, R13. At the resonant frequency the inductive and capacitive reactances are equal. Therefore, the current I_S is in phase with E_S . This condition is illustrated in figure 24(3). Note that $0.5 E_1$ (half of the voltage E_1 across capacitors C14 and C15) lags the current by 90° , and $-0.5 E_1$ leads the current by 90° . The vector addition in figure 24(3) shows the magnitudes of E_2 and E_3 to be equal. At a frequency below 60 megacycles, the capacitive reactance is larger than the inductive reactance, and therefore the current I_S leads the voltage E_S . This condition is illustrated in figure 24(4). Note that the magnitude of E_2 is now greater than E_3 . If the input frequency is above 60 megacycles, the inductive reactance becomes the larger and the current less than E_S .

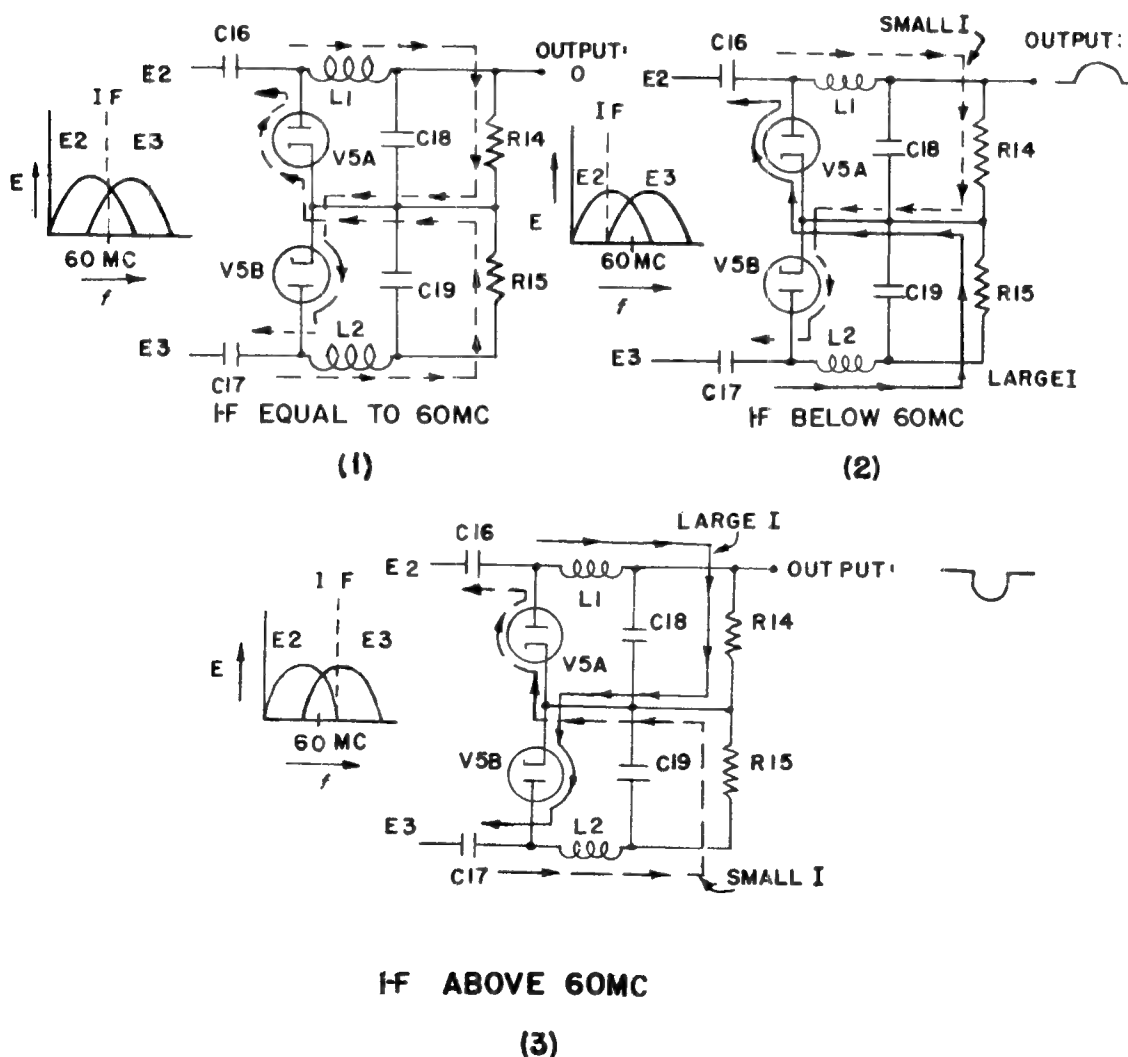


Figure 25. Discriminator detector.

- (4) To make use of the variations in the voltages E2 and E3, they are applied to a detector circuit through capacitors C16 and C17. Figure 25 shows the currents in the circuit and the output from the circuit under various operating conditions. The voltages E2 and E3 are effectively applied across tubes V5A and V5B, respectively. The rectified voltages across capacitors C18 and C19 are proportional to the voltages E2 and E3 and are of opposing polarity, as indicated in figure 24. The output from the detector is the algebraic sum of these two rectified voltages. Figure 25(1) shows the currents for an intermediate frequency of 60 megacycles. The currents through the diodes are equal and opposite, and the voltages across capacitors C18 and C19 exactly cancel. Therefore, at 60 megacycles, the output from the circuit is zero. If the intermediate frequency is below 60 megacycles, as shown in figure 25(2), the current through V5A is larger than that through V5B, and the voltage across C19 is larger than the voltage across C18. Thus, a positive voltage pulse appears at the output. Figure 25(3) shows the currents and the negative output voltage which exists when the intermediate frequency is above 60 megacycles.
- (5) Coils L1 and L2 in the detector circuit are r-f chokes which keep the i-f currents out of the detector circuit. Note that the current flow through each coil is always in the same direction and is, therefore, direct current. The secondary winding of transformer T5 is slug-tuned to the proper frequency. Variable capacitor C40 tunes the primary circuit of T5.
- (6) Capacitor C19 is smaller than C18 to compensate for wiring and tube capacities to ground, which are in parallel with C19. This shunt-type detector is used here instead of the more conventional series detector to obtain a low impedance source and to obtain a circuit that is not critical to tube changes. It is, therefore, possible to replace tube V5 without detuning the discriminator. The output of the discriminator will be a negative 1.3-microsecond pulse when the intermediate frequency is above 60 megacycles and a positive pulse below 60 megacycles. At 60 megacycles, there will be no output. The output may be observed at TP1.

57. VIDEO AMPLIFIER V6A

The purpose of the video amplifier is to amplify the small video output pulses from the discriminator so that they may operate the pulse stretcher. Tube V6A is a conventional video amplifier. A small amount of degeneration is introduced by the unbypassed cathode resistor, R16. This helps maintain the pulse shape.

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The output is developed across the plate load resistor, R52. C2 is the decoupling capacitor for V6A. The pulses from V6A are coupled through C20, developed across R32, and are applied to the center tap of the secondary of transformer T6.

58. CATHODE FOLLOWER V6B

The positive 30-volt gate pulses obtained from the cathode circuit of the acquisition magnetron are brought into the AFC channel through J2. The pulses have ragged tops which are clipped by crystal CR1. The cathode terminal of CR1 is at a positive 26-volt potential, being connected to voltage divider R60-R62. When the signal at the grid of V6B tries to exceed 26 volts, the crystal conducts, and the resulting current through R24 holds the signal at a maximum amplitude of 26 volts. Resistor R25 matches the impedance of the coaxial cable at J3. The voltage at the junction of R60 and R62 is maintained at a constant value by capacitors C39 and C24. Tube V6B operates with a 5-volt, fixed-cathode bias obtained from voltage divider R61-R27. The positive 20-volt output pulses are taken off the cathode, coupled through C36, developed across R30, and applied to the primary of transformer T6.

59. PULSE STRETCHER V7A AND V7B

a. General. The pulse stretcher is a device which converts the pulse output of the discriminator into a d-c voltage. The pulse stretcher is essentially a capacitor which charges quickly to the polarity of the applied pulse but discharges very slowly. The d-c output from the pulse stretcher is used to control a balanced modulator.

b. Operation.

- (1) The relationship between the primary and the secondaries of transformer T6 is such that the positive gate pulse applied to the primary of T6 develops a positive pulse at the V7A end and a negative pulse at V7B end of the secondary. In the absence of a pulse from the discriminator, equal currents flow in each half of V7 during the time the gate pulse is applied, and no voltage is developed across C23. The resistor-capacitor combinations in the plate circuit of V7A and the cathode circuit of V7B develop bias as a result of the gate pulse current. This bias keeps the diodes cut off during the interval between pulses. This feature is incorporated to eliminate any effects of random signals and interference and to guarantee that the steady d-c voltage which is developed across C23 is a function only of the polarity and amplitude of the pulses received from video amplifier V6A.

- (2) The pulses from the video amplifier are applied to the center tap of the secondary of T6 coincident with the application of the gate pulses to the primary. The two pulses are added in the secondary, and, since the gate pulse is inverted in the upper half, the effect is to reduce the pulse voltage at one end of the secondary and increase it at the other. Assume the discriminator pulse to be two volts positive. At the upper end of the secondary, the gate pulse is negative. When combined, the two pulses result in a negative pulse of 18 volts at the cathode of V7B. At the lower end the two positive pulses add and result in a 22-volt pulse at the plate of V7A. Unequal conduction takes place in the two diodes, and capacitor C23 is charged positively through V7A. Since the diodes do not conduct between pulses and since the discharge path for C23 is through R31, which is almost 5 megohms, C23 is charged rapidly to a positive d-c level which is maintained essentially constant as long as the discriminator input pulses remain unchanged. As the amplitude of the error pulses changes, so will the d-c level. In the case of negative error pulses, the output d-c level is negative. The d-c level output is applied to the 400-cycle balanced modulator.

60. AUTOSEARCH PROVISIONS

a. General. When the intermediate frequency varies more than 5 megacycles above or below the correct 60-megacycle frequency, the i-f signal can no longer be passed by the i-f amplifier stages of the AFC. As a result, no discriminator action is possible, and the system must be capable of automatically causing the local oscillator to search through the frequency band. The system must also be capable of reverting automatically to discriminator control when the proper local oscillator frequency is attained, but must not revert to discriminator control when a 60-megacycle signal appears as a result of the local oscillator being tuned 60 megacycles below the transmitter frequency. Automatic control of this action is the function of tubes V11A and V11B and the associated search voltage circuits. Automatic search tube V11 is a dual triode. Its operation is controlled by the grid leak bias developed by limiter V4. When the AFC channel is locked on the proper frequency, this bias causes tubes V11A and V11B to hold relay K1 energized. As long as K1 is energized, the center tap of the secondary of T6 is returned to ground through resistor R32 and the contacts of K1 and the discriminator controls the AFC system. However, if the intermediate frequency becomes so far removed from 60 megacycles that it can no longer be amplified by the i-f amplifiers in the AFC channel, the limiter stage will not develop the bias voltage, and relay K1 becomes deenergized. When K1 is deenergized, the search voltage circuit applies a d-c voltage to the center tap on the secondary of T6. The d-c voltage causes the receiver tuner motor to sweep the local oscillator through its frequency range. When the local oscillator approaches the frequency which is 60 megacycles above the magnetron frequency, the limiter again develops the grid

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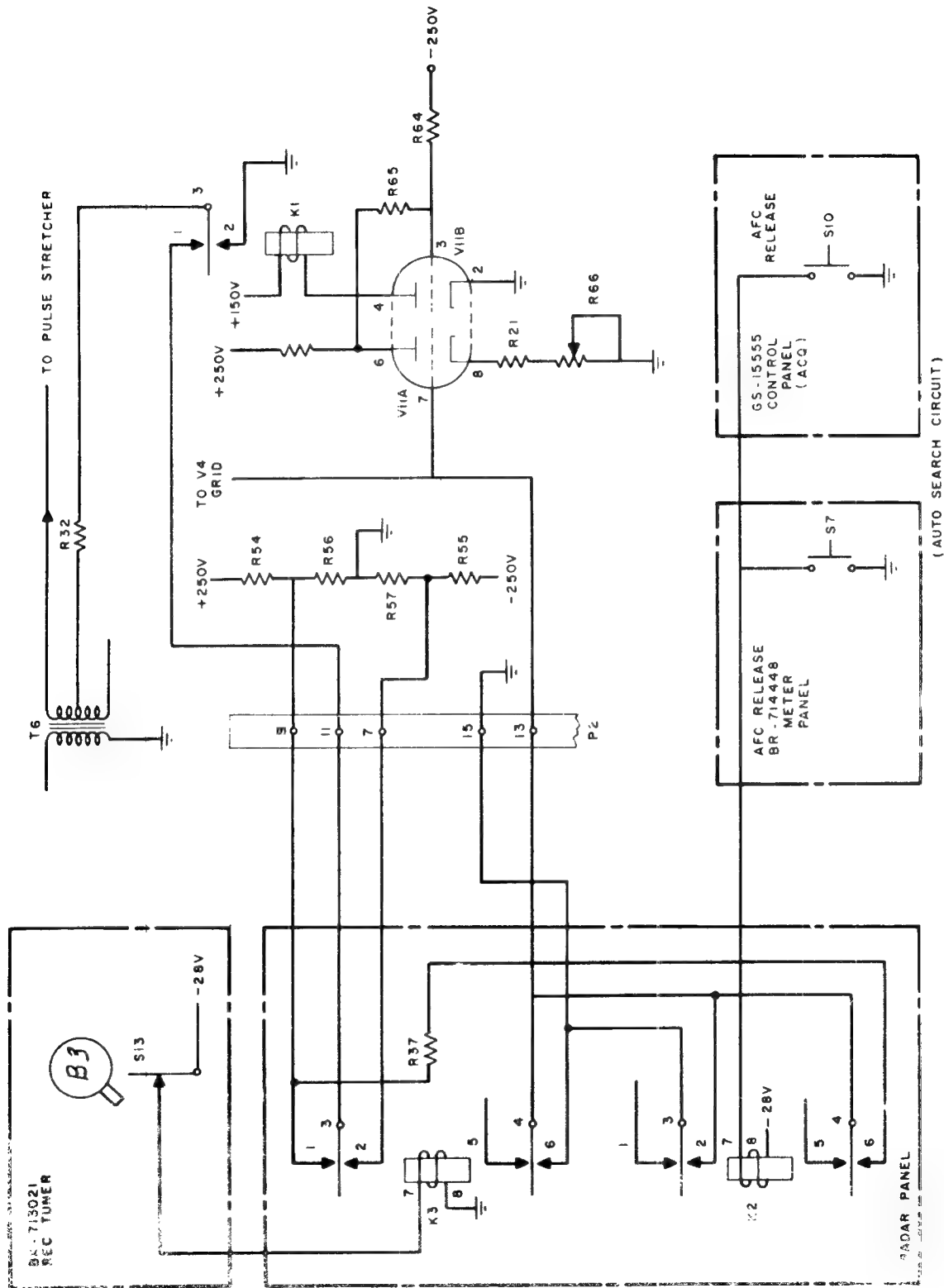


Figure 26. Automatic search circuit.

leak bias which causes tube V11 to energize relay K1. The search voltage is then removed and the discriminator resumes control of the system. The automatic search circuit prevents the local oscillator from locking on a frequency 60 megacycles below the magnetron frequency. The system may jitter near the frequency 60 megacycles below the magnetron, but can be driven through this frequency by the operation of either of the two AFC RELEASE pushbuttons.

b. Normal operation. Figure 26 is a simplified schematic of the automatic search circuits showing the position of the switches and relay contacts when the system is locked on the proper frequency. Bias from limiter tube V4 holds tube V11A near cutoff. Tube V11B is conducting heavily, and relay K1 is energized. K3 is energized and K2 is deenergized. Assume that during normal operation the magnetron frequency jumps down 200 megacycles. The bias of V11A disappears and V11A conducts harder. Plate current of V11A flows through resistor R53, lowering the voltage at the grid of V11B causing V11B to conduct less. Relay K1 is deenergized, and its clapper makes contact with terminal 1. A positive 5-volt potential obtained from the voltage divider R54, R56, and R37 is applied through the contacts of K3 and K1 to the center tap of the secondary to T6. This positive voltage acts as an artificial discriminator signal and appears as an input to the balanced modulator. The modulator produces a 400-cycle voltage that is of such phase as to run the receiver tuner motor in a direction that will decrease the frequency of the local oscillator. As the difference frequency approaches 60 megacycles, intermediate frequency signals appear at the input to the AFC i-f amplifiers. These signals cause the limiter to develop a negative grid leak bias voltage, which is applied to the grid of V11A. V11A conducts less, causing its plate voltage to rise. A rise also occurs at the grid of V11B. Conduction through V11B energizes K1, moving the clapper back to contact 2. This restores control of the system to the discriminator channel. As long as the magnetron or local oscillator frequencies do not jump more than 5 megacycles, the AFC system will vary the local oscillator frequency to keep it exactly 60 megacycles above the magnetron.

c. Partial search cycle. Assume that during normal operation the magnetron shifts 130 megacycles higher in frequency. The i-f input to the AFC channel disappears and with it the bias of V11A. V11A conducts heavier causing V11B to conduct less. K1 becomes deenergized, and the positive 5-volt potential is applied to T6. The system is driven down in frequency to the lower limit of the frequency range. When the lower limit is reached, reversing switch S13 on the receiver tuner is opened, deenergizing K3. When K3 is deenergized, it performs two functions. It replaces the positive 5-volt potential with a negative 15-volt potential obtained from the voltage divider R55 and R57. This negative 15 volts reverses the direction of rotation of the tuning motor. K3 also grounds the grid of V11A through contacts 4 and 6. The tuning motor now drives the local oscillator to the upper limit of the frequency range. The grounded grid renders V11A insensitive

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to the bias developed by limiter V4 as the local oscillator passes upward through the frequencies that are 60 megacycles below and 60 megacycles above the magnetron frequency. The negative 15 volts applied to T6 overrides any discriminator signal that may be developed. When the system reaches the upper limit of the frequency range, reversing switch S13 closes, energizing relay K3. K3 removes ground from the grid of V11A and reapplies the positive 5 volts the center tap of the secondary of T6. The tuning motor is again reversed and starts decreasing the local oscillator frequency. The system will now repeat the process described in b above.

d. Complete search cycle. Although the system is prevented from locking on either lower or upper sideband while it is being swept from the lower to the upper limit of the frequency range, two conditions can exist which will cause the system to attempt to lock the local oscillator at the frequency 60 megacycles below the magnetron frequency. If, when the equipment is turned on, the local oscillator is tuned to develop an intermediate frequency that is less than 55 megacycles per second, jittering near the lower sideband will take place and will also occur if, during normal operation, the magnetron jumps upward in frequency (relative to the local oscillator frequency) more than 5 megacycles but less than 120 megacycles. Under either of these conditions, ground is not applied to the grid of V11A. Then, as the local oscillator tuning approaches the lower sideband, limiter bias decreases the conduction of V11A and throws control of the system to the discriminator channel. At the lower sideband, however, the frequency-polarity relationship is reversed, and the tuning motor drives the local oscillator back up in frequency. When this happens limiter bias appears, and automatic search again takes over, driving the frequency down again. The system again approaches the lower sideband, and the bouncing or jittering sequence is repeated. To get the system through the lower sideband under these conditions, one of the AFC RELEASE pushbuttons must be operated. Operation of one of these controls energizes relay K2. Contacts 2 and 3 of K2 ground the grid of V11A, and contacts 4 and 6 remove resistor R37 from the voltage divider circuit. With R37 out of the circuit, a positive 15-volt potential is developed instead of a 5-volt potential. V11A conducts heavier, V11B conducts less, and relay K1 applies the positive 15 volts to the center tap of T6. The positive 15 volts overrides any signals from the discriminator and permits the local oscillator to tune through the lower sideband. Automatic search control then tunes the system to the lower frequency limit, back to the top of the frequency range, and then down to the frequency which is 60 megacycles above the magnetron frequency, at which time control through the action of the discriminator channel is reestablished.

e. Adjustment. Resistor R66, the relay amplifier adjustment on the AFC chassis is very critical in its adjustment if the AFC is to perform properly. The purpose of R66 is to adjust the conduction of V11A, which in turn adjusts the

point at which relay K1 operates. If R66 is too far clockwise, too much resistance is in the cathode circuit, and, as a result, K1 is energized at all times. This prevents the AFC from auto searching. If R66 is too far counterclockwise, the result is that K1 is always deenergized, and the AFC searches continuously. Correct adjustment is between these two conditions. To adjust R66:

- (1) Disconnect J1 at the AFC chassis. Establish normal magnetron current.
- (2) Turn R66, the relay amplifier adjustment, fully clockwise.
- (3) Slowly turn R66 counterclockwise until the receiver tuner motor starts turning fast. Stop turning R66 at the point where the motor starts.
- (4) Reconnect J1 and see if the AFC locks on correctly and then push the AFC release pushbutton to see if the AFC will search and lock on correctly.
- (5) Slight readjustment on R66 will correct failures in step 4.

Section IV. RECEIVER TUNING SERVO CIRCUITS

61. BALANCED MODULATOR (TM 9-5000-26, page 214)

a. General. The fact that the preselector cavity and the local oscillator are tuned by an a-c motor makes it necessary that an a-c voltage be developed to drive the motor. This is accomplished by use of the balanced modulator. The variable d-c voltage from the pulse stretcher determines the phase and amplitude of this a-c voltage. Tubes V8 and V9 are dual triodes. The four halves are connected in a balanced bridge circuit. Two inputs are applied to the circuit: the variable d-c control voltage from the pulse stretcher and a continuous 400-cycle sine wave. The circuit is designed so that, in the absence of any d-c control voltage, the bridge will be balanced and no output will be obtained from the circuit. When a d-c control voltage is applied, an unbalance is created, and a 400-cycle, push-pull output voltage is developed. The phase and amplitude of this output voltage is a function of the polarity and magnitude of the d-c control voltage from the pulse stretcher. The a-c voltage output from the modulator determines the amount and direction of rotation of receiver-tuner motor B3. Dual triode tubes V8 and V9 are connected so that they operate as an electronic switch. A 400-cycle, a-c voltage is applied through a transformer to the cathodes of these two tubes. The positive or negative d-c control voltage received from the pulse stretcher controls the switching action. The modulator stage produces two outputs 180° out of phase with each other. Each of these outputs contains harmonics, the signal having been distorted by nonlinear amplification. Each

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output experiences a 180° phase shift whenever the d-c control voltage changes polarity. Both outputs are applied to the amplifier phase inverter V10, where these two outputs are combined into a single signal and the distortion is reduced.

b. Operation. A 400-cycle voltage is applied across the bridge by transformer T7, the secondary of which is center tapped to the negative 250-volt supply. The primary of T7 is energized by 60-volt, 400-cycle tachometer excitation (tach X). The 2 A-sections of the tubes operate with a common plate load resistor, as do the 2 B-sections. The 2 sections of V8 operate with a common cathode resistor, as do the 2 sections of V9. The d-c control voltage from the pulse stretcher is applied to the grids of V8A and V9B. The grids of the other 2 sections, V8B and V9A, are returned to the d-c potential present at the brush arm of MOD BAL control R43. The MOD BAL control compensates for slight differences in the conduction of opposing sections of the circuit. As any unbalance may as easily be in one direction as in the other, it is apparent that the d-c potential required at the brush arm may in one case be positive and in another case negative. For that reason, R43 is connected in a voltage divider between plus 250 volts and minus 250 volts so that ± 0.5 volts may be tapped off. The control is adjusted so that in the absence of any d-c control voltage input there will be no output from the circuit. When the d-c control voltage received from the pulse stretcher is zero, the circuit does not produce an output. However, when it is not zero, a 400-cycle output signal is produced. The phase of this output signal is determined by the polarity of the d-c control voltage, and the amplitude is determined by the magnitude of the d-c control voltage.

c. Operation with no error signal. The 400-cycle voltages at opposing ends of the center tapped secondary of T7 are 180° out of phase. The signal at the common cathode of V8 is therefore 180° out of phase with the signal present at the common cathode of V9. With no d-c control voltage input, the conduction through the 4 tube sections balances. The signal which tends to appear at the plate of V8A as a result of the conduction of V8A is in phase with the signal present at its cathode. Likewise, the signal which tends to appear at the plate of V9A because of the conduction of V9A is in phase with the signal present at the cathode of V9A. As the signal at the cathode of V8A is 180° out of phase with the signal at the cathode of V9A, the signal which tends to appear at the plate of V8A is 180° out of phase with the signal which tends to appear at the plate of V9A. The plate of V8A is tied to the plate of V9A, however, with no error signal applied these plate signals are of equal amplitude, and cancellation takes place. At this same time, the B-sections operate in a similar fashion, and the 400-cycle signals are canceled in their plate circuit. This action may be considered from another viewpoint. Resistor R39 is the common plate load resistor for the two B-sections. The voltage output obtained at the plates of the B-sections is determined by the current flow through R39; this current flow is contributed partly by V8B. At that time the signal at the cathode of V9B passes

through its positive alternation, and current flow through V9B is decreased. The increase in current through V8B is equal to the decrease in current through V9B. The total current flow through plate load resistor R39 does not change, and the output voltage remains constant.

d. Operation with a positive d-c control voltage. The 4 sections of the modulator do not operate on the linear portions of their characteristic curves, but operate instead at the knee of the curve. Because of this action, the gain of a section increases when the d-c grid potential is made more positive and decreases when the grid voltage is made more negative. Application of a positive control voltage to the grids of V8A and V9B, therefore, causes the gain of those 2 sections to increase. A greater average current will flow through V8A and V9B, developing a higher average voltage across the cathode resistors. The more positive cathode potentials act to increase the bias and hence to reduce the gain of the other 2 sections, V8B and V9A. The gain of V8A is now greater than the gain of V9A, and the gain of V9B is greater than the gain of V8B. The signal appearing at the plate of V8A as a result of the conduction of V8A will now be of greater amplitude. The signal tending to appear at the plate of V9A as a result of the conduction of V9A will be of reduced amplitude. Cancellation will not take place. Instead, a signal will be present in the plate circuit of the A-sections which will be in phase with the 400-cycle voltage present at the cathodes of V8. In a similar manner an output signal will be present at the plate circuit of the B-sections. Because of the increased gain of V9B and the reduced gain of V8B, this signal will be in phase with the 400-cycle voltage present at the cathodes of V9. The 2 output voltages are therefore push-pull signals, 180° out of phase.

e. Operation with a negative d-c control voltage. Application of a negative d-c signal from the pulse stretcher to the control grids of V8A and V9B will reduce the gain of those 2 sections. In a manner similar to that explained in the preceding paragraph, the gain of V9A and V8B will be made to increase. An output signal will appear in the plate circuit of the A-sections which will be in phase with the 400-cycle voltage present at the cathodes of V9 and an output signal of opposite polarity will appear in the plate circuit of the B-sections. Note that the output signals obtained as a result of a negative d-c control voltage are opposite in phase to those obtained as a result of a positive d-c control voltage.

f. Balanced modulator adjustment. The balanced adjustment R43 may be correctly set by:

- (1) Removing J1, the input jack on the AFC chassis, to put the motor into auto search.
- (2) Depress the MOD BAL button S1.

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(3) Adjust R43, the MOD BAL control, till the receiver-tuner motor comes to rest but not at its upper or lower limit.

(4) Reconnect J1.

62. CATHODE FOLLOWER AND INVERTER V10

a. General. This stage employs a dual triode tube. Two inputs are applied to obtain a single output with reduced distortion. The push-pull 400-cycle output of the modulator provides the input signals. The signal applied to the grid of the A-section is 180° out of phase with the signal applied to the grid of the B-section. Cathode coupling is employed in order to apply to the cathode of the B-section a signal of the same phase and approximate amplitude as the signal applied to the grid of the A-section. As a result, the B-section actually receives 2 inputs, 1 to its grid and 1 to its cathode. These 2 inputs are of opposite phase. The output of the amplifier phase inverter is taken off the plate of the B-section. This signal will tend to be in phase with the signal applied to the cathode and 180° out of phase with the signal applied to the grid. It is apparent, then, that the 400-cycle error signal at the plate will be greater in amplitude than either input to V10, and will agree in phase with the input applied to the A-section.

b. Elimination of distortion. Harmonics present in the push-pull output of the modulator are in phase at the two grids of V10. The harmonics applied to the grid of the A-section appear at the cathode of the B-section in phase with the harmonics present at the grid of the B-section. The degenerative effect is sufficient to eliminate almost entirely the harmonics present in the signals applied to V10. Capacitor C33, connected between the plate of V10B and ground, provides a shunt for the further elimination of harmonics. The reactance of this capacitor at the desired 400-cycle frequency is approximately 40,000 ohms. However, at the higher frequencies characteristic of the harmonics, the reactance is considerably less. A larger value of capacitance would function more efficiently as a filter, but the resulting reduction in reactance at 400 cycles would cause undesirable attenuation of the error signal itself. Capacitor C32 couples the 400-cycle error signal to the mixing circuit composed of equal resistor R47 and R48. Tachometer feedback is applied across developing resistor R49 to the same mixing circuit. The resultant voltage at the junction of R47 and R48 is the output of the circuit. This signal is sent to the low-power servoamplifier.

63. LOW-POWER SERVOAMPLIFIER (TM 9-5000-26 page 33)

a. General. The purpose of the low-power servoamplifier is to amplify the small 400-cycle error signal received from the 400-cycle balanced modulator to a power level which is sufficient to operate the servomotor. A high degree of

amplification renders the servo system extremely sensitive to small error signals. This amplifier consists of 4 tubes, voltage amplifier V1, paraphase amplifier V2, and power amplifiers V3 and V4. It has a gain of 50,000 and a power output of 10 watts maximum.

b. Voltage amplifiers V1A and V1B. The output of the 400-cycle balanced modulator is applied directly to the grid of V1A. Tube sections V1A and V1B are connected as conventional R-C coupled voltage amplifiers. Bias is developed at V1A by voltage divider R2-R3-R8, connected between the negative 250-volt supply and ground. R8 also serves as the cathode resistor for V1B. Bias is developed at V1B by the quiescent plate voltage of V1A and the voltage divider R1-R5-R4, connected from plus 250 volts to minus 250 volts. R1 also serves as the plate load resistor of V1A. Plate resistor R5 of V1B operates as part of a bridged T-filter to be explained later. C1 is a coupling capacitor between the 2 stages and operates to couple to the grid of V1B instantaneously and at full value any voltage change which occurs at the plate of V1A. Negative feedback is taken from the cathode of V1B and is applied to the grid of V1A. The purpose of the negative feedback is to overcome any tendency of this high-gain amplifier to oscillate. It accomplishes this purpose by resisting any change in current flow through the two amplifiers. During the no-signal period, current flow through V1A is approximately 0.5 ma, and the current flow through the two stages must be considered in determining the magnitude of the various operating voltages. The effect of the negative-current feedback is to resist any tendency for current to vary through V1A or V1B. Should the current flow through V1B tend to increase, the cathode would become more positive, and this voltage rise would be coupled to the grid of V1A. A voltage increase at the grid of V1A would cause a voltage drop at the plate. This voltage drop would be coupled to the grid of V1B through C1 and would act to reduce the current through V1B to the original value. A decrease in current through V1B would be corrected in a similar manner. In attempting to maintain a constant current, the amplifier will present an apparent high-output impedance. The amplification possible with this type of circuit is not great because of the negative feedback employed. Amplification is sacrificed in favor of stable operating characteristics. The output of V1B is coupled through C3 to the grid of V2A of the paraphase amplifier.

c. Amplifier and phase inverter V2. Tubes V2A and V2B form a conventional 2-tube, cathode-coupled, paraphase amplifier. Tube V2A operates as an inverter amplifier. Tube V2B is used as an amplifier to produce a signal of the same amplitude as the output of V2A, but of opposite polarity. Since common cathode resistor R10 is not bypassed, the voltage which appears across it is the resultant of the 2 plate currents, and has the same shape and polarity as the voltage applied to the grid of V2A. This action is similar to that which takes place in a cathode follower, but it differs in 2 important respects. The output from the stage is

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taken from the plate so that although the voltage developed across R10 is degenerative the gain is not limited to less than unity, and the plate current of both tubes flows through resistor R10. The value of R10 has been selected so that the amplitude of the degenerative voltage developed across it is equal to half the voltage applied to the grid of the A-section. The effective signal voltage which appears between grid and cathode of the A-section is half the applied voltage. The signal voltage which appears at the cathode is also equal to half of the applied voltage. As the grid of V2B is connected to ground, application of this voltage to the cathode has the same effect as would be obtained by applying a voltage of opposite polarity to the grid. As a result, signals opposite in polarity but equal in amplitude are applied between grid and cathode of the 2 sections, and a balanced push-pull output signal is obtained.

d. Power amplifier V3 and V4. The push-pull 400-cycle output signal of V2 is applied to the power amplifier. This stage consists of two pentode tubes which are operated as class AB. Bias is developed across the common-cathode resistor and plate circuits. Regeneration of undesired harmonics at frequencies above 400 cycles is prevented by the R-C circuit consisting of R20 and C8. The impedance of this circuit is high at 400 cycles but decreases sharply at higher frequencies. This reduction in impedance introduces increasing attenuation at the frequencies of the higher harmonics. The output is developed across transformer T1 by currents flowing alternately through each tube and through each half of the primary of T1. As both tubes operate class AB, each tube conducts for slightly more than half of each cycle. For a part of each half cycle, therefore, bucking currents flow in the primary of T1. Appreciable power loss as a result of these bucking currents is avoided by use of a plate supply voltage, which is the unfiltered output of a full-wave rectifier. This supply voltage is obtained from the plus 270-volt supply. The peak voltage amplitude attained each half cycle is +420 volts. The average voltage (peak voltage times $2/\pi$) is 270 volts. The a-c input to the plus 270-volt power supply is in phase with the 400-cycle voltage applied to the balanced modulator. Therefore as the plate supply voltage for tubes V3 and V4 reaches its peak of plus 420 volts, the error signal present at one grid is at its maximum positive point and at the other grid is at its maximum negative point. At this instant, maximum power is delivered by the stage. At an instant when the push-pull error signals at the control grids are at or near zero values, both tubes are enabled to conduct. At this time, however, the plate supply voltage is at or near zero. The amplitude of opposing currents which may flow in the primary of T1 is therefore very small, and power loss is minimized. The current flow in the primary of T1 is produced by the alternate conduction of 2 tubes, and a 400-cycle error signal is developed across the secondary of the transformer. As this output signal will be applied across an inductive load consisting of the control-field winding of the servomotor, capacitance must be added to the circuit to avoid a phase shift. This phase shift correction is made by C6 and C7. Maximum power output of the stage is about 10 watts.

e. Bridged-T network. Undesirable frequencies, the most pronounced of which is the third harmonic, are introduced into the system principally through the feedback voltage from the tachometer. A conventional negative-feedback circuit which would introduce sufficient degeneration to eliminate these frequencies would result in serious attenuation of the desired 400-cycle error signal. The bridged-T network, however, is a frequency selective circuit. The components of the bridged-T network are C2, R7, C3, and R6. The values of these elements have been selected so that negative feedback is attained only at frequencies above or below 400 cycles, and minimum feedback will occur for signals at 400 cycles. The desired amplitude of feedback voltage is obtained by means of the voltage divider consisting of resistors R21 and R22. From the junction of these two resistors, the feedback signal is applied to the bridged-T network at the junction of C2 and R7. A possible path for the signal is through R7. However, at 400 cycles, the impedance of C2 is considerably less than that of R7. The signal takes the path of lower impedance, appearing then at the junction of C3 and R6. At 400 cycles, the impedance of R6 is less than that of C3, and the 400-cycle feedback signal is prevented from appearing at the grid of V2A. Instead, it is shunted to ground through the power supply. A feedback signal at a frequency considerably above or considerably below 400 cycles, however, would be applied to the control grid. A feedback signal at a frequency of 4,000 cycles would, at the junction of C2 and R7, see a lower impedance through the capacitor than through the resistor. As a result, the signal would next appear at the junction of C3 and R6. Again, the capacitive path would provide lower impedance, and the feedback signal would be applied to the control grid. The gain of the lower power servoamplifier at any frequency is proportional to the attenuation experienced by that frequency in the bridged-T network.

64. SERVOMOTOR (TM 9-5000-26, pages 196, 197)

a. General. The servomotor is powered with 2 a-c voltages 90° out of phase with each other and consists of a stationary field structure and a rotating element. The motor is of the squirrel-cage type, so named from the resemblance of the rotor to a squirrel cage. The rotor consists of copper bars welded to end rings and imbedded in iron laminations.

b. Servomotor B3. Motor excitation is continuously applied to the fixed-field winding of the receiver-tuner servomotor. When the intermediate frequency of the AFC deviates from 60 megacycles, an output signal appears at the output of the low-power servoamplifier and is applied to the control field winding of the servomotor. It may be 90° or 270° out of phase with motor X. The direction of rotation of the servomotor is determined by which of these two phase relationships exists. The direction of motor rotation will be such as to tune the local oscillator to the frequency 60 megacycles above the transmitter frequency. When

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the intermediate frequency is returned to 60 megacycles, the 400-cycle error signal disappears from the output of the low-power servoamplifier, and the servomotor comes to rest.

65. TACHOMETER

The purpose of the tachometer is to furnish a feedback voltage which is proportional to speed. This feedback prevents hunting of the servo system and also aids in braking the motor. The tachometer is an a-c generator built on the induction principle. It is physically contained within the servomotor housing, and is driven by the rotor of the servomotor. It is used because of its low inertia, low power demand, and ease of maintenance. It provides a feedback voltage to the input of the associated low-power servoamplifier. Eddy currents which are excited in a hollow cylindrical rotor induce a voltage in a secondary winding. Eddy currents are currents produced by an emf induced in a conductor when the conductor is moved in a magnetic field or when a magnetic field in which the conductor is located changes in intensity. These currents are usually small. The induced voltage is at right angles to the magnetic lines of force. The stator of the tachometer has two windings displaced 90° from each other. One of the windings is the input winding, and the other is the secondary output winding. A core of soft iron is centered in the space which would be occupied by the rotor of a conventional generator. Sufficient space is left between the stator and the core to permit the rotation of a hollow cylindrical rotor. The input coil is excited by tachometer excitation. The a-c current creates an alternating magnetic flux which is directed and confined by the soft-iron core. As long as the rotor is stationary, no output voltage is developed because the secondary winding is perpendicular to the magnetic field created by the current in the input coil. However, eddy currents are produced in the rotor cup because of the changing magnetic field. The rotor surface may be considered as an infinite number of shorted conductors such as those in a squirrel-cage rotor. When the rotor is turned, these conductors cut the magnetic flux created by the current in the input winding and current is caused to flow in the rotor. This current in the rotor generates a second magnetic field which distorts the original field to produce a resultant field. The magnitude and direction of the resultant field are dependent upon the speed and the direction of rotation of the rotor. In an induction motor, the rotating stator field drags the rotor conductor. In a tachometer, the rotor conductors may be considered as dragging the stator field. For this reason the rotor is called a drag cup. When the stator field is dragged or distorted, the angle existing between the resultant field and the output winding is no longer 90° , and, consequently, the lines of flux cut the output winding and induce in it a voltage with a phase and amplitude dependent upon the direction and speed of rotor rotation. When the servomotor reverses direction, the rotor of the tachometer also reverses direction. The output voltage then undergoes a 180° phase reversal. The output voltage is always degenerative at the point where it

is introduced at the input to the low-power servoamplifier. One disadvantage of this type of generator is that the eddy currents generated in the aluminum drag cup cause harmonics to appear in the output voltage. These harmonics, if not eliminated, would cause power loss and excessive heating of the servomotor. It is important that the output voltage be exactly 180° out of phase with the output signal of the 400-cycle balanced modulator. The design of the tachometer is such that the phase of the generated output voltage is accurate within a 1° tolerance. A variable resistor is placed in series with the energized stator coil. Adjustment of this control will produce sufficient phase shift to obtain an output of the exact phase required. This resistor is adjusted by ordnance and should not be altered. The output of the tachometer provides speed-feedback damping of the servomotor. As the error signal disappears, tachometer-feedback serves as a braking signal to stop the servomotor.

66. OVER-ALL OPERATION (TM 9-5000-26, pages 196, 197)

The over-all operation of this AFC channel may be determined at any time by observing either of the neon hunt indicators. For example, the neon hunt indicator on the acquisition receiver control shows what the AFC channel is doing. The hunt indicators are neon lamps which light whenever the output of the low-power servoamplifier exceeds 40 volts. When the system is locked on the proper frequency, these lights blink very rapidly. When the system is jittering near the frequency 60 megacycles below the magnetron frequency, the lights blink slowly to indicate this condition. When the system is searching, the lights glow steadily and blink off about once each 10 seconds indicating the reversal of tuner motor B3. If the lamp glows steadily at all times, the cause is either that the average magnetron current is not 30 ma or more, which is required to apply motor X to the motor, or that one of the two AFC motor switches is in the off position. Since these two neon lamps are directly across the output of the servoamplifier, their failure to light indicates that there is no output from the AFC.

Section V. STC, MTI, AND SWITCHER-MIXER

67. SENSITIVITY TIME CONTROL (STC) CHANNEL

a. General. The purpose of the STC channel is to reduce the gain of the receiver at close ranges so that all return signals will have a more nearly equal intensity. The STC chassis is located on the back of the acquisition control panel. It operates to produce a negative-going waveform that can be applied to the control grids of V3, V4, and V5 in the i-f preamplifier channel for selective control of the preamplifier channel gain during each transmitter cycle. Echoes at very close ranges tend to be very strong; and if they are not attenuated, a blossoming which interferes with target detection will occur around the center of the PPI.

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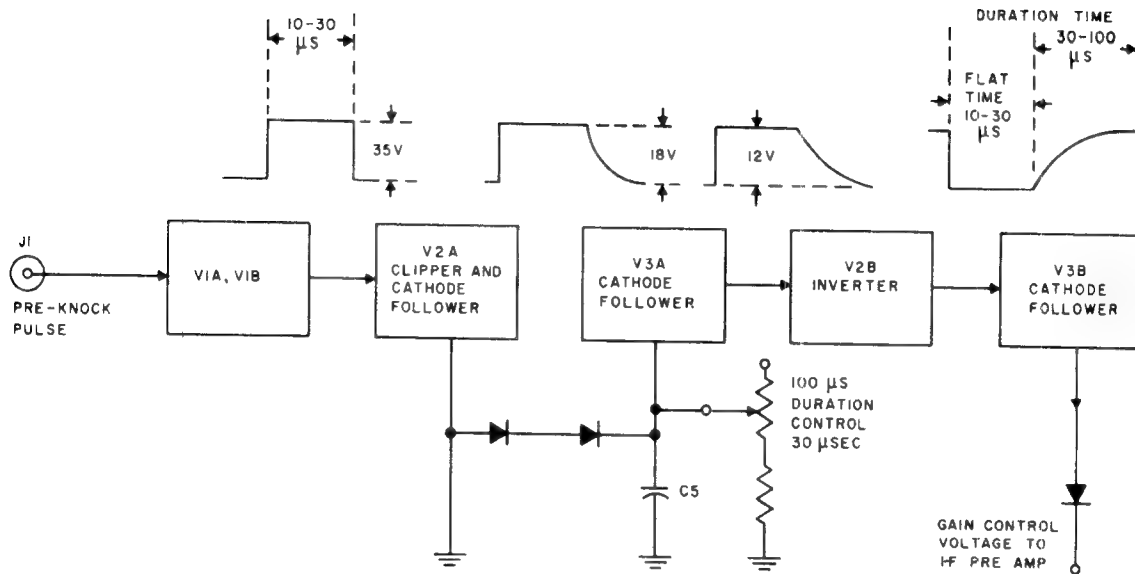


Figure 27. Sensitivity time control.

b. Multivibrator V1 operation (TM 9-5000-26, page 213). The sensitivity time control channel circuits are triggered by the positive preknock pulse, which enters the unit through jack J1. Dual triode V1 is a cathode-coupled, one-shot multivibrator. V1B is normally conducting because its grid is returned to +150 volts. The voltage developed across cathode resistor R5 is high enough to hold V1A cut off. When the preknock pulse is applied to the grid, the A-section is triggered into conduction. The plate voltage drop is applied through C3 to the grid of V1B, driving it toward cutoff. Cathode follower action causes the cathode to drop, reinforcing the increase of current through the A-section. This process reaches temporary equilibrium when the B-section is completely cut off and V1A reaches maximum conduction. The grid of V1B then begins to rise slowly toward +150 volts because of the exponential charging of C3 through resistor R9. When it rises just above cutoff, the B-section begins to conduct again. The cathode voltage also becomes more positive, reducing current flow through the first section. The consequent rise in plate voltage, coupled by C3, aids the rising grid voltage in V1B. V1B rapidly reverts to its pretrigger state of conduction. A positive, rectangular output pulse is taken from the junction of R7 and R8 in the plate circuit of V1B and applied to the grid of the next stage. The duration of the output pulse is variable between about 10 and 30 microseconds, the exact

duration being determined by the setting of potentiometer R2 (labeled FLAT). R2 sets the bias level of the grid of the first section. The bias governs the maximum excursion of the plate voltage; and hence, the potential to which the second-section grid drops, and finally, the time required for the grid to again reach cutoff. R2 is normally set to obtain a 24-microsecond output, which is the equivalent of approximately 4,000 yards.

c. V2 and V3A operation. The positive pulse output at the cathode of follower V2A rapidly charges capacitor C5 through crystals CR1 and CR4. When the pulse ends, however, the crystals prevent the discharge of C5 through the relatively low resistance of R11. Instead, C5 must discharge through R13 (labeled DURATION) and R14. The waveform at the grid of V3A has a steep front, a flat top for a time between 10 and 30 microseconds, and a sloping, trailing edge resulting from the decay of voltage at the top of C5. Potentiometer R13 can vary the time constant of the discharge path between about 12 and 1,000 microseconds. R13 is normally adjusted so that the trailing edge of the waveform extends for 92 microseconds or 15,000 yards. V3A is another cathode follower. Its output is applied to the V2B grid, which is clamped positively to -9 volts by crystal CR2. V2B is a low-gain amplifier with degenerative cathode feedback for preservation of the waveform. The plate load for this tube is the STC potentiometer R6, located on the receiver control unit chassis. Tube V2B cuts off at -15 volts. The -9v clamping level on the grid holds the tube slightly above cutoff.

d. V3B. The negative-going waveform from V2B is applied to cathode follower V3B. The grid of this tube is held at a high positive potential, about 30 volts, to insure linear reproduction of the entire waveform. The grid circuit, as in the case of the other cathode followers, contains a small series resistance for the suppression of transient parasitic oscillations. Crystal diode CR3 clamps the baseline of the output waveform to a d-c level that can be set to some value between 0 and -20 volts. The level is established by adjustment of GAIN control potentiometer R5, located on the acquisition control panel. The output of the sensitivity time control unit is a negative square wave with a sloping, trailing edge riding on the gain control voltage set by R5 in the receiver control unit. Thus, the gain control voltage is a composite signal.

68. MOVING TARGET INDICATOR SYSTEM

a. General. In order to designate a target to the radar operators it is first necessary to see the target on the PPI. If the target flies over an area in which clutter is present, the return from the clutter will frequently obscure the return from the target. Accurate target designation then becomes impossible. The purpose of the MTI system is to reduce the intensity of clutter on the PPI so that moving targets will be visible when they are in an area in which clutter

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is present. Common causes of clutter are clouds, precipitation, and fixed targets such as hills, trees, and buildings. Clutter may also appear when enemy targets use certain types of countermeasures. Under these circumstances, the MTI system can assist in observation and accurate designation of targets which would be totally obscured without the MTI system.

b. Interference phenomena. A variety of colors can be seen on the edges of a very thin oil slick. This phenomenon arises from interference between light reflected from the upper surface and light reflected from the lower surface of the oil. Although the oil slick is much too thin to measure by ordinary means, its thickness is appreciable compared to the wavelength of light. The variations in colors on the oil slick indicate differences in its thickness. When a target is flying in the presence of clutter, the antenna receives echoes from the target and from the clutter. Interference occurs between these echoes just as interference occurs between light reflected from the two surfaces of the oil slick. The MTI uses this interference effect to determine whether the target is moving or stationary. The oil slick is too thin to measure with a ruler, but its thickness could be computed from the known wavelength of light. Similarly, it is quite difficult to measure directly that a target is moving. But by using the interference effect, it can be deduced that a target in the presence of clutter is moving. For this MTI system to work, not only is a moving target needed, but also clutter.

c. Radar application. How does this effect work in radar? If there are two or more targets at approximately equal azimuth in range, the radar receiver cannot discriminate between the two signals as separate targets. To the receiver, the returned signal would appear as the echo from a single target and have an amplitude (intensity) equal to the vector sum of the amplitude of the echoes from each of the individual targets. The amplitude of the vector sum depends upon the phase relationship of the received echoes. The phase relationship, in turn, is determined by the round-trip path length between the antenna and the radar reflecting surface. If this path-length difference remains constant, the amplitude detected by the radar receiver remains constant. If one of the radar reflecting surfaces is moving, then the path-length difference is constantly changing and each pulse echo detected by the receiver will have a different amplitude. This is important. If two successive received signals from the same range and azimuth are compared, it can be determined whether they are of equal amplitude. If they are, it is recognized that the source of these echoes is clutter, and the signals are rejected. On the other hand, if the amplitude of the signals is not constant, it is recognized that the echoes originate from a combination of several reflecting surfaces, one of which is moving. These signals are not rejected but are applied to the PPI.

d. Effect of target speed. How fast must the target move before its movement can be detected? Since it is detected by means of an interference effect, the minimum discernible movement of the target in the clutter is derived in terms of wavelengths. Its exact value depends upon the amplitude of reflected signals from both target and clutter. However, a typical minimum value would be a pulse-to-pulse change of one-tenth wavelength in round-trip path length. At the S-band frequencies used in the acquisition radar, this corresponds to a radial target velocity of only 10 mph. Target speed therefore, is no problem in the design of a moving target indicator for antiaircraft radars. The problem is how to match accurately two successive echoes from the same area to determine whether or not they are of equal amplitude. The radar information is divided into two parts. One part is sent through a delay channel, the other part is sent through nondelay channel. The delay channel stores this information for a period of time precisely equal to the time interval between successive transmitted pulses from the radar. The delay channel then reproduces these pulses with very nearly identical waveshape and amplitude, but of opposite polarity, to the output of the nondelay channel. These two outputs are then added together, the algebraic sum is amplified, and the negative portions inverted. If, when the outputs of the delay and nondelay channels are combined, the two signals are equal and opposite, the algebraic sum is zero. If the amplitudes are not equal, the algebraic sum will equal the difference between the amplitudes of two successive returns from the same target range. This difference is known as the residue. Ideally the residues from fixed targets and from nonfading targets in the clear would be zero and these returns would not appear on the PPI. The residues from targets in clutter would not equal zero. The residue is amplified and applied to the PPI to indicate a target echo.

69. LIMITATIONS IN THE SYSTEM

a. Circuit limitations. In any realizable system there are never ideal conditions. The residue from fixed targets is small, but not zero. For detection of moving targets, the residue from moving targets must then be larger than the residue from fixed targets or from system noise (noise is random and never cancels). In order to evaluate the effectiveness of MTI performance, a figure of merit known as the cancellation ratio is often used. Cancellation ratio is defined as the ratio of the amplitude of a fixed target echo (with the nondelay channel disconnected to prevent mixing with the delay channel output) to the amplitude of the residue from the fixed target echo with the outputs of the delay and nondelay channels mixed and algebraically summed. In other words, it is the factor by which fixed targets are suppressed by the MTI system. The cancellation ratio is expressed in decibels (db). This ratio is defined by:

$$\text{cancellation ratio in db} = 20 \log_{10} \frac{\text{amplitude of uncanceled pulse in volts}}{\text{amplitude of canceled residue in volts}}.$$

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To simplify the testing procedure, a constant-amplitude test pulse is provided in the system to simulate a target echo. This test pulse permits cancellation ratio measurement without the use of actual targets.

b. Antenna limitations. The discussion here will consider only a few of the more important limitations to achieving a high cancellation ratio. The principal limitation is not a function of the MTI circuitry, but a basic characteristic of scanning antennas. Antenna scan inherently limits the cancellation ratio. This is due to the variation in gain across the nose of the antenna's beam. As the beam sweeps across the target, the gain changes slightly between each two successive transmitted pulses. The pulse-to-pulse change in gain causes a difference in received echo amplitude and a cancellation residue proportional to this gain change. The amount of this residue depends upon the pulse repetition rate, the antenna beam width, and the rotational speed of the antenna. For the acquisition system, an antenna rotational rate of 10 rpm limits the over-all cancellation ratio to 23 decibels; at 20 rpm effective cancellation decreases to 17 decibels; and at 30 rpm antenna scan so limits cancellation ratio that MTI is of negligible value and should not be used. The highest obtainable cancellation ratio is therefore 23 decibels. The remaining parts of the MTI apparatus is to be able to provide a cancellation ratio sufficiently better than the limitation imposed by the antenna scan, so that the antenna, rather than the electronic circuitry, is the limiting factor.

70. BLOCK DIAGRAM DISCUSSION (TM 9-5000-26, page 207)

a. MTI simplified block. The MTI system works on a delay principle. Received video is compared in successive periods, using a 15-megacycle carrier modulated with one of the outputs of the i-f amplifier channel in the MTI modulator channel on a 15-mc carrier. The carrier is necessary because the quartz delay medium in the MTI delay channel would not reproduce video pulses correctly. This single frequency is modulated with the video, a preknock pulse, and a test pulse when needed. The modulated output of this modulator channel is then sent to a delay channel and a nondelay channel. The delay channel retards the carrier containing preknock and video for one pulse repetition period or 1,000 microseconds. The nondelay channel however, passes the signals undelayed so that with a reversal of polarity, the nondelayed and the delayed video can now be compared or added in the MTI video channel. Fixed echoes now cancel out because there is no change between each fixed echo while moving targets are uncanceled. A residue video (moving target) is passed through the MTI video channel to the switcher mixer channel to be displayed on the indicators if desired. Another output of the delay channel is the auto sync pulse. This pulse is simply the preknock pulse which has been delayed 1,000 microseconds and which will be sent through the auto sync channel and used to retrigger the synchronizer and cause it to operate in a synchronous manner.

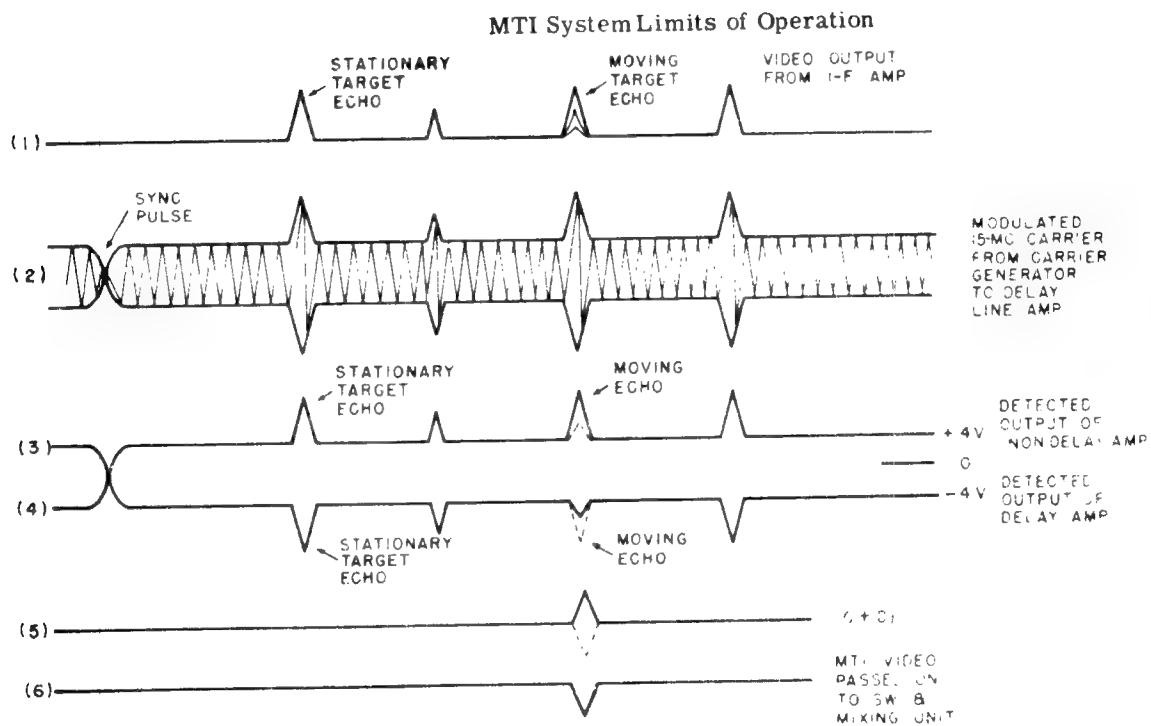
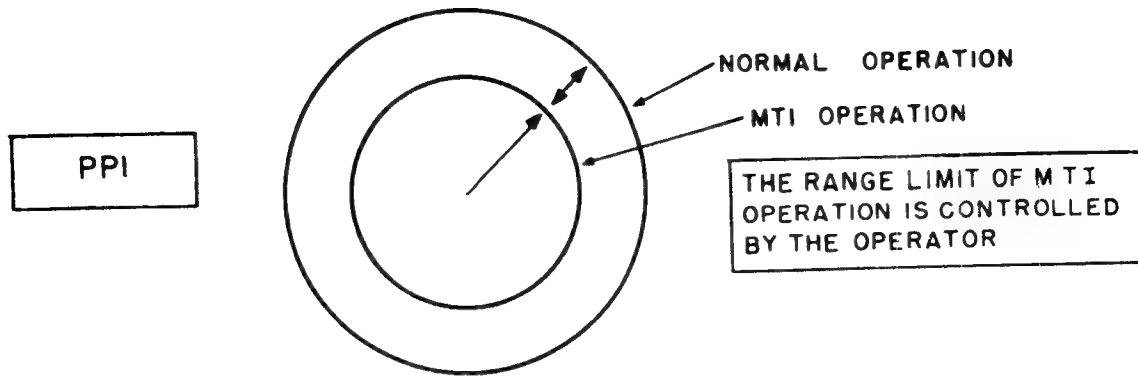
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Figure 28. MTI waveforms.

b. MTI system detailed block diagram. The positive video output from the i-f amplifier (fig 28(1)) channel is used to modulate a 15-mc carrier generated by oscillator V2 in the MTI modulator channel. In addition, a positive preknock pulse from the synchronizer is amplified and inverted in the inverter amplifier V1 of the MTI modulator channel and applied to the 15-mc carrier in modulator V3 (MTI modulator channel). The use to which the preknock is put will be discussed later. The modulated carrier is amplified in the r-f amplifier V4 and amplifier V1 (MTI modulator channel) and applied to the MTI delay channel (fig 28(2)) and the MTI nondelay channel. In the MTI delay channel, electrical energy is converted to sonic energy in the delay cell and the modulated 15-mc

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supersonic signals pass through the cell, being delayed by an interval equal to the time between two successive magnetron pulses (about 1,000 microseconds). At the termination of the delay cell the supersonic pulses are converted back to electrical energy and fed to amplifier V3 (MTI delay channel). The output of V3 is sent to a cathode follower V2 and from there to an attenuator. This attenuator is used to adjust the gain of the delay channel. From the attenuator the signal is sent through amplifiers V1-4 and power amplifier V5. The signal then is detected by CR1 and CR2 and then fed through delay line Z1 to the MTI video channel (fig 28(4)). The input signal to the MTI nondelay channel is the same signal that was applied to the MTI delay channel except that it has undergone 41 decibels of attenuation. The modulated 15-mc signal is amplified and detected (fig 28(3)) in the MTI nondelay channel and applied to the MTI video channel. The gain through the delay and nondelay channels is controlled by the action of two d-c amplifiers, V2 and V8 respectively, the delay d-c amplifier V2 operating from the output of the delay channel, and the nondelay d-c amplifier V8 operating from any residual voltage at the junction of R25 and R26 in the MTI video channel. The nondelay d-c amplifier acts to hold this point at a d-c zero level. The crystal detectors at the output of the nondelay and delay channels are polarized so that the nondelay channel produces positive video pulses (fig 28(3)) at a +4-volt level, and the delay channel negative video pulses (fig 28(4)) at a -4-volt level. While the timing of the entire radar system is arranged to give a pulse repetition rate such that the interval between any two pulses is almost exactly equal to the delay time of the quartz cell, a small amount of circuit delay is inevitable. This is compensated for in the adjustable delay network Z1/A29 which can be set in 0.05-microsecond steps to secure coincidence in time for the equal and opposite fixed-target pulses at the junction of R25 and R26. Since exact timing is of greatest importance for MTI operation and since delay cell characteristics vary with temperature changes, it is almost impossible to design a free-running pulse generator that can maintain a pulse rate having a constant time difference between pulses. For this reason the delay period of the quartz delay line itself is used to establish the pulse repetition rate and consequent time relationships throughout the entire acquisition system. A portion of the output from the delay channel, containing the positive preknock pulse and the negative video information, is applied to amplifier V3 (auto sync channel). This tube is operated at cutoff and only the preknock pulse is passed. It is then taken through another stage of amplification V4 (auto sync channel) and through a cathode follower V5 (auto sync channel). The resulting positive pulse is amplified in V1 (auto sync channel) and used in the synchronizer to initiate the next preknock and sync pulses. In this manner the timing of the system is tied in directly with the delay characteristic of a particular quartz delay cell. If the delay time of the cell drifts with temperature changes, the pulse repetition rate is automatically adjusted to match the change. Returning to the junction of R25 and R26 (MTI video channel), where the outputs of the delay and nondelay channels are added algebraically, if the amplitude and time relationships have

been properly balanced through the preceding circuits, the stationary target echoes and preknock pulses through the two channels (figs 28(e), (f)) will cancel. The moving target echoes, however, having different amplitudes from pulse to pulse, will not cancel completely. The small resultant signal, which can be either negative or positive, is amplified in V6. To insure that only a negative video output is obtained from the MTI video channel, the signal is put through a crystal rectifier CR1, polarized to pass only negative pulses. Positive pulses are inverted in V7. Crystal CR2 passes only negative pulses. The negative outputs from both crystals are combined in one channel which leads to the switcher-mixer channel.

71. MTI MODULATOR CHANNEL

a. General (TM 9-5000-26, page 215). The period of delay that is necessary for MTI operation cannot be satisfactorily achieved through electrical networks. Therefore, in this equipment quartz supersonic delay cell is used. Since the 1-microsecond video pulses cannot by themselves be accurately reproduced by the delay-cell crystal transducers, the video information must be used to modulate an r-f signal whose frequency is the same as the resonant frequency of the crystal transducers. The selected frequency is 15 megacycles.

b. Oscillator V2. The oscillator which generates the 15-mc signal is a 6AK5 pentode connected as a triode in a Hartley circuit. The oscillator is tuned to 15 megacycles by factory adjustment of capacitor C4. This setting is not critical since the 15-mc quartz crystals in the delay cell are sufficiently damped to pass an off-resonance carrier with little more than normal attenuation. The oscillator output is taken from the cathode of V2 and applied to the control grid of modulator tube V3.

c. Inverter amplifier V1. The positive preknock pulse enters the MTI modulator channel through J3 and is applied to the control grid of V1. This tube is a low-gain amplifier inverter. The unbypassed cathode resistor R16 provides degenerative feedback to maintain the high-frequency response of the tube, hence, the pulse shape. The negative pulse output of the preknock amplifier is then applied to the suppressor grid of V3.

d. Controls. The amplitude of the positive video pulses from the i-f amplifier channel is controlled by potentiometer R13, which sets the level of modulation in V3. When the test pulse switch S2 is in OFF position (off for test pulse), the video information is passed to the suppressor grid of V3. Capacitor C8 balances the succeeding circuits.

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e. Modulator V3. The control grid of V3 is biased at -4.3 volts, thus limiting the amplification of the 15-mc carrier. Grid No. 3 (pin 7), which injects the preknock and video pulses into the carrier, is biased through voltage divider R22-R23 to about -7 volts. The preceding conditions of operation combine to produce a 100 percent modulated 15-mc signal at the output of V3 for maximum amplitude video signals when potentiometer R13 is approximately set.

f. R-F amplifier V4. The r-f amplifier V4 is broadly tuned to 15 megacycles by Z2 in the grid circuit and Z3 to provide a good match to the coaxial line which carries the signal to amplifier V1, in the delay line amplifier (TM 9-5000-26, page 216).

g. Amplifier V1. The parallel plate circuit consisting of L1 and R3 provides a wide frequency response. R5 and R6 introduce a 41-db attenuation of the output signal sent to the nondelay channel. An additional 24-db attenuation of this same signal takes place at the input to the nondelay channel. Hence, the signal is attenuated 65 decibels before being applied to the first stage of the non-delay channel. The unattenuated output of V1 is applied to delay cell Z1 in the MTI delay channel where it is delayed 1,000 microseconds.

72. TEST PULSE CHANNEL

When the TEST PULSE switch on the carrier generator chassis is placed in the ON position +250 volts is applied to plate V3. The A-section of V3 receives as its input the sync pulse; it amplifies this signal and applies it to V3B, the test-pulse blocking oscillator to act as a trigger. V3B when triggered conducts for 7 microseconds producing a 6-volt signal across its cathode resistor R18. The output test pulse is applied to the suppressor of modulator V3 (MTI modulator channel) through the TEST PULSE switch.

73. MTI DELAY CHANNEL

a. General. The MTI delay channel receives the modulated 15-mc carrier from amplifier V1 in the MTI modulator channel. In the delay channel, the signal is delayed for 1,000 microseconds, the 15-mc carrier is removed, and negative video signals appear at the output. These negative video signals riding at a -4-volt level are applied to the MTI video channel where they are compared to the output of the nondelay channel output signals.

b. Delay cell. The MTI delay cell is a small polygon disk with about 15 carefully ground plane surfaces. A crystal transducer is vibrated by application of the modulated 15-megacycle carrier. The mechanical vibrations travel through the quartz disk and strike a polished face. They are then reflected to a second face and then to a third. This action continues until the

reflected vibrations reach the last face 1,000 microseconds later. A second crystal transducer then reconverts the mechanical vibrations to a 15-megacycle voltage. The delay cell is mounted on the chassis of the delay cell amplifier.

c. Stages V2 and V3. V3 is a conventional r-f amplifier. Its output is applied to cathode follower V2. Cathode follower V2 is a dual triode with its plates held at a-c ground by the decoupling action of C7. The function of V2 is impedance matching. Cathode resistance R22 is paralleled by R23 and the resistance of the attenuator (in the zero position, the attenuator resistance is zero). The attenuator provides from 0 to 20 decibels of attenuation in 5-db steps. This attenuator is adjusted in conjunction with the GAIN PET adjustment and will be covered later.

d. Amplifiers and detectors (TM 9-5000-26, page 217). The output from the attenuator is fed to the amplifiers V1-4 and power amplifier V5. The first 4 tubes, V1 through V4, are conventional voltage amplifiers. The fifth stage V5 is a power amplifier. The second through the fifth stages are broadly tuned to 15 megacycles by reactances Z1 through Z4 in the grid circuits. The plate circuit of V5 is coupled through transformer T1 to the detectors CR1 and CR2. The detectors are germanium crystals connected in push-pull so the full-wave detection takes place with accurate reproduction of the video pulses. Capacitor C18 is charged to a negative d-c level whose value is almost equal to the amplitude of the 15-mc carrier. The negative-going video pulses are riding on a d-c level which is normally -4 volts. The actual value of the d-c level is a function of the gain through the amplifiers which is controlled by the gain through amplifier V1. Control bias for V1 is derived from d-c amplifier V2 which, in turn, operates from the reference level of the delay channel output. If the signal amplitude is such that a d-c level greater than -4 volts is obtained at the output, the d-c amplifier increases the bias at V1, thus decreasing the gain and the output voltage.

e. Automatic gain control amplifier V2 and V1A (TM 9-5000-26, page 219). When the two sets of signals are added algebraically with the junction of R25 and R26, two conditions must be met for effective MTI operation. Coincidence in time must be approached as closely as possible; and the amplitudes of the negative fixed-target video pulses from the delay channel must be equal to the amplitudes of the coincident positive fixed-target pulses from the nondelay channel. To obtain the latter condition, the amplification through each channel must be accurately controlled. The d-c amplifier V2 provides feedback gain control. The output of the detectors CR1 and CR2 is applied through R8 to the grid (pin 3) of V2. Capacitor C1A is a virtual short circuit to ground for the pulse component of the signal. Only the d-c reference voltage, therefore, is applied to the grid. V2 is a dual triode with a common cathode resistor R7. The first section of the

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tube is essentially a cathode follower, and changes in voltage at the grid (pin 3) will result in like changes at the top of R7, which are simultaneously injected into the second section of the tube through cathode (pin 8). The GAIN SET potentiometer R3 adjusts the bias on the grid (pin 7) between the limits 0 and -6 volts, determining the degree to which small changes at the cathode will affect conduction through the second section. The setting of R3 ultimately determines the value of the d-c reference voltage at the delay amplifier output. Assume that the d-c level at pin 3 of V2 is -4 volts and that R3 is set so that the circuit will operate to maintain the -4-volt level. Now, if the voltage at the pin 3 grid decreases by some small amount, that is, moves toward 0, current through the first section increases. The drop across R7 increases, making both cathodes more positive than they were. In the second section, this has the effect of making the grid more negative with respect to the cathode, and current decreases. This causes the plate (pin 6) to rise, decreasing the negative bias level on amplifier V1. The gain through the amplifier V1 is increased until a new equilibrium at the -4-volt level is established. The analysis may be carried through similarly for the case where the d-c level increases in a negative direction. A connection is made through R9 to the MTI meter panel on the radar cabinet framework, which is used when the d-c level is set to the desired value, nominally -4 volts. The gain-set control R3 is first turned to maximum, and the MTI attenuator is adjusted for a reading of about 50. Then R3 is reset to the 40 reading. Clamper V1A prevents the bias voltage applied to amplifier V1 from rising above ground. This limits the amplification possible through the amplifier and prevents the distortion that would result from saturation.

f. Delay line Z1. Any pulse through the delay channel will emerge at the output of the detectors of the pulse from the nondelay channel against which it is to be matched. When the nondelay portion of an initial preknock pulse applied to the MTI modulator channel appears at the nondelay channel output, the delay portion has just begun its journey down the quartz delay line. One thousand microseconds later the delayed pulse appears at the output of the delay channel, but the pulse against which it is to be matched has not yet appeared. The reason for this is that it has not yet been initiated by the synchronizer. The preknock pulse through the delay channel, which is used to trigger the synchronizer for each successive cycle, must undergo a slight additional delay through the circuits of V3, V4, V5 and V1 in the video auto sync channel and in the external loop. To compensate for this additional delay, Z1 is inserted in the delay channel preceding the balancing point. The network is an LC lumped-constant type of line having a minimum delay of 0.95 microseconds and a maximum delay of 1.25 microseconds and adjustable in increments of 0.05 microsecond. It is set to delay the delay channel pulses the additional amount necessary to bring them into coincidence with the nondelay channel pulses. The two signals are then applied through resistors R25 and R26 to a common point. R25 and R26 introduce some attenuation, but are necessary to isolate the two

channels from each other, thus eliminating reflection distraction. Algebraic addition of the two signals takes place at the junction of R25 and R26. If the preceding circuits are operating properly, the 2 d-c levels and coincident fixed target pulses are equal in magnitude and opposite in polarity. Moving target pulses, however, are usually not identical in amplitude from cycle to cycle. While cancellation of fixed target pulses, then, tends to be complete, addition of the moving target pulses results in small residual signals which are passed on the following circuits. Since a moving target pulse of greater amplitude may come through the delay channel on one occasion and through the nondelay channel on another, the residual signal pulse may be either positive or negative.

74. MTI OPERATION

a. General. One of the requirements for effective MTI operation is a constant pulse repetition rate, that is, one with a constant time difference between successive pulses. In the acquisition radar system, the preknock pulse delayed through the quartz delay cell is used to trigger the synchronizing circuits for the generation of the next pulse. In this manner, the quartz delay cell, in addition to functioning integrally in MTI operation, serves as the timing device for the entire radar system. If the delay characteristic of the quartz cell changes with temperature, the system pulse rate is automatically adjusted to accommodate the change in delay time, and MTI operation is preserved.

b. Operation (TM 9-5000-26, page 219). The amplifier chain consisting of V3, V4, V5 and V1 prepares the delayed preknock pulse for transmission to the synchronizer. In any particular cycle of video information, the preknock pulse appears at the delay channel amplifier output as a positive going pulse followed by the negative going target pulses. This information is applied through capacitor C2 to the control grid of tube V3. The tube is biased at cutoff, approximately -6 volts, by voltage divider action across R11 and R12. The negative pulses are eliminated and only the positive preknock pulse is amplified through the tube. The output of V3 is applied to the control grid of V4 for inversion and further amplification. Generated originally as a rectangular pulse, the preknock pulse at this point in the circuit has acquired a slight leading-edge slope. This fact makes possible small adjustments of the system timing. TIME BALANCE control R15 can be continuously set to provide different values of load resistance, allowing selection over a limited range of the time the pulse reaches triggering amplitude. This control is needed to establish a finer time balance than is possible from strapping the delay network Z1, which is adjustable only in 0.05-microsecond steps. The range of control afforded by time balance control R15 is about 0.01 microsecond. The positive pulse output of V4 is applied to V5, which is a dual triode cathode follower circuit. The two sections of the tube are connected in parallel because the power requirements for developing an output pulse of sufficient amplitude are greater than one triode section

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can meet. The grids are biased below cutoff so that only pulses of sufficient amplitude cause conduction, and negative pulses which may have passed through V3 are eliminated completely. Resistors R19 and R23 act to suppress spurious oscillations which may result from shock-excitation of naturally resonant portions of the circuit. The positive output pulse is taken from the top of resistor R22 and connected to the autosync pulse Jack J1. A coaxial cable then routes the transformed preknock pulse to amplifier V1 (TM 9-5000-26, page 198) which amplifies this autosync pulse to a large enough amplitude to trigger the blocking oscillator V2 of the synchronizer. The AUTO-INTERNAL switch at the grid of V1 is normally left in AUTO except during certain parts of the MTI adjustment.

75. NONDELAY CHANNEL (TM 9-5000-26, pages 216, 218)

The signal for the nondelay channel is taken from the 41-db attenuator at the output amplifier V1 (MTI modulator channel). At the input to the nondelay channel the signal is put through a 75-ohm matching pad which attenuates it another 24 decibels. At this point in the circuit, the signal is identical within limits of minor amplitude variations to the output of the delay channel, except that the latter has been delayed by almost one whole pulse repetition period. In the nondelay channel the crystal detectors are polarized in the opposite direction from those in the delay channel. Capacitor C18 is charged positively, and the detected positive video signals are referred to the +4-volt level established by the charge on C18. The gain of the nondelay channel is controlled by a bias developed in d-c amplifier V8 and applied to the grid of V1. The control signal for the d-c amplifier is not taken from the output of the nondelay channel, but from the junction of R25 and R26 where the outputs of the two amplifiers are added. The gain control loop for the nondelay channel functions to keep this point at a zero d-c level.

76. DELAY CHANNEL (TM 9-5000-26, page 219)

The amplitude of fixed target pulses is roughly set by the action of the delay channel d-c amplifier V2, but the amplitude of fixed target pulses from the nondelay channel must be adjusted to an equal amplitude so that maximum cancellation will take place. Since a fixed relationship between pulse amplitude and the d-c level derived from the operation of circuits of the MTI modulator channel is the same in the delay and nondelay channels, it is reasonable to assume that if the negative d-c level of the delay channel cancels the positive d-c level output of the nondelay channel, the fixed-target video pulses will be equal and opposite in amplitude and will also cancel. The gain control system of the nondelay amplifier is based on this assumption. Any residual voltage at the junction of R25 and R26 operates through d-c amplifier V8 to adjust the gain of the nondelay channel in the direction necessary to produce a zero junction

voltage. Since the gain control circuits must react almost instantaneously to small differences in voltage, V8 is a high-gain, d-c amplifier. The second section of V8 (pins 6, 7, 8) is a cathode follower which controls the sensitivity of the first section. AMPLITUDE BALANCE potentiometer R43 controls the conduction through the second section of the tube and, hence, the steady state voltage at the top of R45B. Since the two cathodes are tied together, the first section is biased by the second section. R43 is adjusted to change the gain through the nondelay channel just enough to bring the R25-R26 junction voltage back to zero when a deviation occurs. Assume that the junction voltage, applied to the grid (pin 3) of V8 through R28, becomes slightly negative. Capacitor C1B filters out the pulse component, leaving only d-c voltage at the grid. The change in the negative direction reduces the current through the first section of V8, causing the plate voltage to rise. The change in plate voltage causes the drop across R39A to change, decreasing the negative bias applied to V1 in the nondelay channel. If R43 has been correctly set, the gain through the nondelay channel increases just enough to bring the R25-R26 junction voltage back to zero. A feedback path exists through resistor R40. A portion of the voltage change at the plate (pin 1) is applied through this resistor to the grid (pin 6). If the change is positive, the effect through the cathode follower action of the second section of the tube is to sharpen the response of the first section. Clamper V1B keeps the bias applied to the nondelay amplifier from rising above ground, preventing distortion in the channel.

77. MTI VIDEO CHANNEL (TM 9-5000-26', page 219)

The residual moving-target pulse from the junction of R25 and R26 is applied through capacitor C8 to the control grid of V6, a medium-gain amplifier. The unbypassed cathode resistor provides degenerative feedback for preservation of the pulse shape. The circuits involving crystals CR1 and CR2 and tube V7 operate to produce only a negative video output from the MTI video channel. Pulses which were positive, and therefore blocked at CR1, appear at the plate of V7 as negative pulses and are passed by CR2. Negative pulses which at the grid of V7 are inverted and blocked by CR2 and complete negative video information is developed across the MTI video potentiometer R37. The potentiometer arm is connected to video output jack J2. The MTI video is then sent over a coaxial cable to the switcher-mixer channel.

78. SWITCHER-MIXER CHANNEL

a. General. The switcher mixer channel can best be described as an electronic switch which can alternately switch between MTI video and bypass video. This switching action is needed because the MTI output is only needed for the center 35,000 yards of acquisition coverage. The range at which MTI video is displayed is adjustable from 1,000 to 35,000 yards. Beyond that range, the

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switcher-mixer channel allows only bypass video to be displayed. The switcher-mixer chassis is located in the upper right compartment of the radar cabinet. Video signals from 2 sources are applied to the switcher-mixer. One of these inputs is MTI video which is obtained from the MTI video channel; the other is bypass video which is obtained directly from the i-f amplifier channel. Bypass video contains complete information, but fixed object echoes have been eliminated from MTI video. The switcher-mixer channel contains gating circuits which operate to pass on to the video and mark channel either of the 2 types of video. MTI video is passed during the initial part of the sweep. At some point in range, MTI video is shut off, and bypass video is allowed to pass for the remainder of the sweep. When the MTI switch at the acquisition receiver control is in the OFF position, MTI video is shut off completely and bypass video is allowed to pass continuously. The switcher-mixer contains a series of stages to accomplish switching action. Variable width multivibrator V1 determines the point of switchover. The tubes V3 and V5 act as gate tubes, and stage V6 acts as a mixer amplifier. Stage V2 simply inverts the output of multivibrator V1.

b. Multivibrator V1. The operation of this circuit is the same as that of the multivibrator in the STC chassis. Potentiometer R2 is the MTI RANGE potentiometer, located on the acquisition receiver control. This potentiometer varies the bias voltage at the grid of V1A. As the bias is varied by this control, the duration of the output waveform of the multivibrator may be varied from 6 to 220 microseconds. The MTI RANGE potentiometer has the same effect upon the switcher-mixer multivibrator as the FLAT TIME potentiometer, R2, has upon the STC channel multivibrator. Switch S1, also located on the acquisition control panel, is the MTI switch. When S1 is closed, the grid of V1A is placed at ground potential, and the multivibrator is made inoperative. The positive waveform which is developed at the plate of V1B is fed through C5 to the control grid of V2 and to the suppressor grid of V5.

c. Inverter V2. This stage provides the means of inverting the multivibrator waveform. V2 is normally cut off by a fixed bias of -18 volts, which is obtained from the voltage divider R23-R24. The waveform experiences little amplification in tube V2. The amplitude of the negative waveform from V2 is determined by the setting of the GATE LEVEL potentiometer, R12. Two considerations are of importance in adjusting this control. First, the amplitude of the negative waveform must be sufficient to cut off V3 completely. Second, since V3 and V5 have a common plate load resistor, R26, it is necessary that V5 be cut off at the same instant that V3 becomes conductive. This is necessary in order to avoid a ring on the PPI's which contains no video signals. Coincidence can be obtained by varying the GATE LEVEL control. The output of V2 is coupled through C3 to the suppressor grid of V3.

d. Bypass gate tube, V3. Video signals from jack J2 are applied through the BYPASS GAIN control, R13, to the control grid of V3. The average amplitude of the bypass video is set by R13 to match that of the MTI video which is fed through J3 to the control grid of V5. The control grid of V3 is biased slightly negative, and the negative input signals are clamped to the bias level by crystal CR1. The crystal overcomes the blocking effect which is brought about by variations in the number and length of video pulses from sweep to sweep and thus holds the tube at a stable operating point. The bias on V3 can be varied over a small range of -1.4 to -2.7 volts by the SWITCH BAL control, R16. This control is adjusted to make the static current of V3 through the common plate load resistor R26 exactly equal to the static current of V5. If this adjustment were not made, the intensity of the display on the PPI would change as the output was switched from MTI to bypass video. At the beginning of each sweep the negative multivibrator gate signal from V2 is applied to the suppressor grid of V3 and cuts the tube off. The suppressor grid of V3 is clamped at ground potential by tube V4B. This insures that the suppressor grid potential during the entire conduction period of V3 will be held at ground level regardless of the width of the gate signal. The plate current of V3 during the conduction period is thus prevented from varying as a result of variations in the suppressor grid voltage.

e. MTI gate tube V5. The operating point of V5 is chosen so that it is possible to adjust V3 to operate identically. The negative MTI video at the control grid of V5 is clamped by CR2 to a negative 2-volt bias level. In the absence of the positive gate waveform at the suppressor grid of V5, this tube is cut off by a negative 18-volt potential obtained from the junction of R23 and R24. The positive gate waveform overcomes the fixed bias, causing V5 to conduct for the same interval that V3 is cut off. The top of the positive gate waveform from the multivibrator is clamped at ground level by diode V4A. This is necessary so that the suppressor grid voltage of V5 during conduction will be the same as the suppressor grid voltage of V3 while V3 is conducting, and thus equal static currents are caused to flow through the tubes. The outputs of V3 and V5, which are positive in polarity, are developed across R26 and coupled through C8 to the grid of V6A.

f. Mixer V6. Only the bypass and MTI video signals or the bypass video signals are applied to V6A. Tube V6 is operated with a bias of about -9 volts, which is suitable for linear amplification. Degeneration due to unbypassed cathode resistor R30 preserves the shape of the video pulses. If an IFF presentation is employed, the B-section of V6 operates to inject the IFF signals into the video signal channel. The plates of V6A are tied together and connected to jack J4. The plate load resistor and the plate voltage source for the tube are located in the video and mark channel.

g. MTI system reference. Complete adjustment details on the MTI system will be found in TM 9-5000-23.

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CHAPTER 6

PRESENTATION SYSTEM

Section I. MIXER CHANNEL

79. INTRODUCTION (TM 9-5000-26, page 226 and TM 9-5000-25, sheet 24)

The purpose of the mixer channel is to combine all the marks that appear on the PPI's and PI's. The mixer channel is a part of the video and mark mixer chassis which is located in the acquisition radar cabinet assembly on the left-hand slide over the power supplies. The mark section of the chassis will be discussed here. This mixer channel consists of an arc coincidence stage V1, a radial-line coincidence stage V2, and amplifiers V3, V4, and V6. The arc coincidence stage combines the TAGA and the QTRMK to form the 10° arc of the electronic cross. The radial-line coincidence stage combines the TRGA and the TAMK to form the 5,000-yard radial line of the electronic cross. The remaining three amplifier stages combine and clip the electronic cross, the range circle, and the flashing steerable azimuth line so that they all appear with the same intensity on the indicators.

80. ARC COINCIDENCE TUBE V1

a. General. When a tube requires the simultaneous application of two or more input signals in order for an output signal to be produced, it is called a coincidence tube. The acquisition tracking range mark is applied to tube V1 at a rate of 1,000 pulses per second. Each time the azimuths of the two radars coincide, the tracking azimuth gate is also applied to tube V1. During a 10° portion of the tracking azimuth gate, tube V1 is permitted to conduct. It produces an output which consists of the acquisition tracking range marks that occur during the 10° portion of the gate. The output of tube V1 consists of negative pulses which produce the arc of the electronic cross.

b. Operation. The V1 is held at suppressor-grid cutoff by the negative d-c level at which the tracking azimuth gate rides. V1 is held at control-grid cutoff by a negative 10-volt potential obtained from voltage dividers R4, R5, and R8. The acquisition tracking range marks are brought into the unit at J1. These signals are coupled through C2, developed across R3, R5, and R8, and applied to the control grid of V1. The amplitude of the acquisition tracking range marks is sufficient to raise the control grid above cutoff; if tube V1 were not held at suppressor-grid cutoff, the acquisition tracking range marks would cause a circle to be traced on the PPI's. The tracking azimuth

gate is a positive waveform 30° wide, superimposed on a varying negative d-c voltage, and is applied directly across the WIDTH potentiometer R2 which is adjusted so that only a 10° portion of the tracking azimuth gate will drive the suppressor grid above cutoff. Each acquisition tracking range mark that appears at the control grid is amplified and inverted by tube V1 and is developed across R7 in the plate circuit. If the acquisition antenna is rotating at 30 rpm, approximately 55 acquisition tracking range marks will occur during the 10° period. The tracking azimuth gate is generated once during each revolution of the acquisition antenna. Thus it can be seen that, for some period during each rotation of the acquisition antenna, tube V1 will allow a series of acquisition tracking range marks to pass. This output will cause the arc of the electronic cross to be produced. The arc is centered at the azimuth of the tracking radar. The purpose of R1 is to provide an impedance match with the coaxial cable at J1. R36 and C3A form a decoupling network. The negative pulses from V1 are coupled through C4 and crystal CR1 to V3. Crystal CR1 isolates tube V1 from the output of tube V2. The polarization of CR1 is such that negative signals applied to terminal 2 of the crystal are allowed to pass, but negative signals appearing at terminal 1 are blocked. Resistor R9 causes the signals applied to CR1 to vary about ground potential.

81. RADIAL-LINE COINCIDENCE TUBE V2

a. General. The tracking range gate, a positive 30-microsecond square wave, is applied to tube V1 once per sweep, or 1,000 times each second. Each time the azimuths of the two radars coincide, the tracking azimuth mark (a positive 735-microsecond square wave) is also applied to tube V2. When both input signals are applied, the tube conducts. This conduction results in the production of a negative 30-microsecond square wave output. This output signal is used to produce the radial line of the electronic cross.

b. Operation. The operation of tube V2 is very similar to that of V1. Tube V2 is held at both suppressor-grid and control-grid cutoff by negative voltages which are obtained from the voltage divider consisting of resistors R12, R14, and R19. The positive 30-microsecond tracking range gate is applied to the control grid V2 once during each sweep. The time of occurrence of the tracking range gate is determined by the range setting of the tracking range system. Each time the azimuths of the acquisition and tracking radars coincide, the positive 735-microsecond tracking azimuth mark appears at the suppressor grid of V2. If tube V2 were not at control-grid cutoff, the tracking azimuth mark would cause one complete PPI sweep to be intensified. At some range during the single sweep when the tracking azimuth mark is applied, tube V2 is driven above control-grid cutoff for a period of 30 microseconds by the application of the tracking range gate. The tube conducts, and a negative 30-microsecond square wave is

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developed across plate load resistor R18. This waveform produces the radial line of the electronic cross. It occurs at the azimuth of the tracking radar and is centered at the range setting of the tracking radar. Resistors R10 and R11 provide the proper terminating resistances for the input cables. R17 is the screen-dropping resistor, and C3B is the screen-bypass capacitor. C9, CR2, and R22 have the same function in stage V2 as do C4, CR1, and R9 in stage V1.

82. CLIPPER AMPLIFIER V3

a. General. The outputs from both V1 and V2 are applied to tube V3. This stage is biased to a value that allows cutoff limiting to be employed. The action of tube V1 causes the negative envelope of the gated acquisition tracking range marks to be V-shaped. Tube V3 limits all but the first few and last few marks to the same amplitude. The synchronization of the various components is such as to cause one of the several acquisition tracking range marks from tube V1 to be superimposed upon the 30-microsecond negative square wave from tube V2. However, clipper V3 removes the superimposed mark, preventing an intensified spot from appearing on the display at the intersection of the two lines of the electronic cross. The output waveforms from tube V3 are positive signals.

b. Operation. Tube V3 is a 6AH6 pentode. This tube has a comparatively small control-grid cutoff potential. The control grid of V3 has a fixed bias of 2.5 volts, which is obtained from voltage divider R23-R26. The inputs of V3 are the negative pulses from V1 and the negative 30-microsecond square wave from V2. One of the pulses from V1 is superimposed on the 30-microsecond square wave. However, the bias of clipper V3 is driven below cutoff by the applied signal, and this superimposed pulse is thereby prevented from appearing in the output. The negative acquisition tracking range marks at the output of V1 vary widely in amplitude. The clipping action of V3 causes all but the first few and last few pulses in each group to have the same amplitude at the output of the stage. The acquisition tracking range marks are comparatively wide at the base of the waveform. Excessive width could cause the arc of the electronic cross to be distorted or widened. To prevent this, the signals applied to V3 are first reduced in amplitude across the voltage divider consisting of R24, R25, and R26, which serves to reduce the width of the pulses appearing at the output of V3. The output of V3 is applied through coupling circuit C11 and resistor R29. Crystal rectifier CR3 provides positive clamping at ground level. Clamping is necessary because the duration of the 30-microsecond waveform present in the output of V3 would be sufficient, in the absence of CR3, to allow the charge on C11 to change appreciably. Crystal CR3 prevents the voltage from going below ground by providing a low-resistance discharge path for C11. The signal, clamped at ground level, is then applied to rectifier CR4 and developed across resistor R30, which is common to rectifiers CR4, CR5, and CR6. Rectifier CR4 isolates V3 from the positive signals introduced at J4 and J5.

83. AMPLIFIER V4

The positive 0.5-microsecond acquisition range marks which produce the range circle are brought into the unit at J4. The positive 735-microsecond acquisition azimuth marks, which produce the flashing steerable azimuth line, are brought in at J5. Resistors R41 and R42 provide proper termination of the cables. Rectifier CR5 provides isolation for the acquisition range channel, and CR6 provides isolation for the acquisition mark generator channel. These elements function in a manner identical to CR4. Resistor R30 is the common load for all three crystal rectifiers. The mixed signals are coupled through C13 and developed across R28, which is returned to a minus 6-volt potential obtained from voltage divider R31-R32. Tube V4 is a conventional amplifier. The grid bias prevents excessive grid-current flow. Improved fidelity results from degeneration introduced by the unbypassed cathode resistor. Decoupling from the supply voltage is accomplished by R33 and C12C. The negative output signals are developed across plate-load resistor R34 and are coupled through C14 to the grid of clipper amplifier V6.

84. VIDEO AND MARK CHANNEL (TM 9-5000-26, page 226)

a. General. The purpose of the video and mark mixer is to combine the received signals from the switcher-mixer channel with the marks from the target designator system (TM 9-5000-25 sheet 24). The output of this channel is applied to the two PPI video amplifiers and the two PI video amplifiers.

b. Video amplifier V5. The negative acquisition video signals from the switcher-mixer are brought in and are coupled through C15, developed across R45 and R46, and applied to the control grid of tube V5. A fixed bias of -1.8 volts is obtained from voltage divider R44-R46. Negative clamping of the signals at the -1.8-volt level is accomplished by CR7. Positive flyback caused by signals of large amplitude and long duration is eliminated by CR7. The positive video signals from tube V5 are applied through resistor R53 to tube V7.

85. CLIPPER AMPLIFIER V6

a. General. The operation of tube V6 is similar to the operation of clipper amplifier V3. Cutoff limiting is used to control the amplitude of the output signals. The operation of the two radars is such that several of the output waveforms from tube V4 may be superimposed on each other. Tube V6 clips any waveforms that are superimposed and thereby prevents blooming on the indicators. The positive output of tube V6 is mixed with the positive video signals from video amplifier V5, and the resultant is applied to cathode follower V7.

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b. Operation. Each acquisition azimuth mark that is applied to V6 has one acquisition range mark superimposed upon it. The operation of the two radars is such that under certain circumstances the 30-microsecond waveform from tube V3 may be superimposed on the acquisition azimuth mark, or the pulses from V3 may be superimposed on the acquisition range marks. Tube V6 limits amplitude of any combination of superimposed waveforms to a value which will not cause blooming on the acquisition indicators. The tube operates with a fixed bias of minus 1.8 volts which is obtained from voltage divider R37-R38. The plate load for the stage is composed of R40 and MARK AMPLITUDE control R50. Potentiometer R50 provides a means of adjusting the amplitude of the signals from V6 so that the items of the display which are produced by these signals will be of the correct intensity. The output signals of V6 are applied through resistor R52 to the control grid of tube V7. In stage V7 the marks are combined with acquisition video which is finally sent to the acquisition indicators.

c. Video and mark mixer. The two halves of tube V7 operate as one cathode follower stage. Two triodes connected in parallel are used to carry the comparatively large amount of current that is necessary to develop signals of the target designator from the input to V7. The voltage divider R56-R58 supplies a fixed bias of -6 volts. Positive clamping of the signals at the -6-volt level is accomplished by CR9. Resistors R57 and R61 suppress parasitic oscillations. The composite output signal of tube V7 is developed across the cathode load resistor, R59, and is applied to J7. From J7, the output is applied to the PPI video amplifiers and precision indicator video amplifiers.

d. Meter M1. Meter M1 is the MTI meter and is used in the adjustment of the MTI channel. This meter is used as a voltmeter to indicate balance adjustments in the MTI and to set the gain of the MTI delay channel. When the MON-BAL switch is in the monitor position, the meter is a direct-reading voltmeter showing 0v to -5v direct current. In the balance position this meter is acting as a galvanometer to show the voltage difference between the cathode of V7 in the video mark mixer and an adjustable voltage divider. The METER SET control allows the meter to be set to a low reading so that when the FINE BAL switch is operated, the meter will read on scale. The fine balance switch simply increases the sensitivity of the meter for balance adjustment. The meter and switches are found on the carrier generator slide immediately above the power supplies in the acquisition radar cabinet. Connection to the meter is through P3-3 on the MTI synchronizer (TM 9-5000-26, page 219). See 44-161-2x, sheet 24, for details.

86. FIELD ADJUSTMENTS

a. With the precision indicator in the track position, the WIDTH adjustment on the video and mark mixer chassis is adjusted until the arc of the electronic cross is 10° or $1/3$ the width of the opening.

b. The MARK AMPLITUDE control is adjusted so that with normal ground clutter visible on the PPI the marks are clearly visible but are not so bright as to obscure weak targets.

Section II. SWEEP GENERATION

87. GENERAL

The display seen on the PPI's (fig 2) can be divided into two sections, the rotating sweep and the video and marks. The basic operation is to generate a rotating sweep which is in step with the antenna rotation and then to intensity modulate this sweep with video and marks. In this section, the sweep generation only will be considered.

88. PRESENTATION SYSTEM (TM 9-5000-26, page 221)

a. Block diagram. The signal which goes to make up the rotating sweep starts initially at the 4-kc oscillator channel. This 4-kc oscillator channel generates three signals, one of which is of interest here. This signal is applied to the azimuth channel, which consists of a resolver and a resolver amplifier chassis, and modulates this 4-kc signal as the antenna rotates. Its outputs are two 4-kc signals, varying in amplitudes as the antenna rotates. These signals are called the north-south(N-S) and east-west(E-W) resolver signals. The N-S signal is a maximum when the antenna is pointed north or south, while the E-W signal is maximum when the antenna is pointed east or west. These signals are sent from the antenna into the trailer and are applied to the N-S and E-W sweep channels of the acquisition radar presentation system. The purpose of these sweep channels is to determine from the resolver signals what direction the antenna is pointing and then to generate a sweep on each scope in the same direction.

b. Generation of a rotating sweep. In the case of the north sweep channel, for example, the resolver signal is first demodulated to obtain a d-c voltage proportional to the north direction or one which is maximum when the antenna is pointing north. This d-c voltage is then used in a sweep generator to determine the amplitude of the sweep voltage generated in the north sweep channel. This sweep voltage is then amplified and applied to the north coil on the PPI. This signal results in a sweep in the north direction when the antenna is pointed north. This same principle is used in the other three sweep channels which also contain individual demodulators, sweep generators, and sweep amplifiers. The signal is maximum in each coil when the antenna is pointed in that coil's direction. For example, when the antenna is west, the west coil receives a maximum signal which would not appear to cause a rotating sweep, but where the antenna is north, the north coil receives a maximum signal and the other three receive

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small ones. This results in a north sweep. When the antenna is at an azimuth of 800 mils, the north and east coils receive equal signals, while the other two receive small signals and the sweep rotates to NE. When the east coil receives a maximum signal and the other three receive small signals, the result is an east sweep. This process is continued as the antenna rotates, and a rotating sweep results.

89. 4-KC OSCILLATOR (TM 9-5000-26, page 222)

a. General. The 4-kc oscillator chassis is located in the target-designate control panel. Its function is to produce a 4-kc signal which is used to excite the acquisition azimuth receiver, one of the line slew resolvers, and to provide a reference carrier voltage. The reference carrier is compared with the N-S and E-W signals from the resolver circuit to determine angular azimuth data. The 4-kc oscillator consists of an oscillator stage V1, push-pull power amplifiers V2 and V3, and a cathode follower V4. This chassis has three outputs, two of which are taken from transformer T4, and the third from stage V4. These output signals are called 4-kc ACQ, 4-kc carrier, and 4-kc line.

b. Oscillator V1. This stage is a push-pull oscillator which generates a c-w signal at 4 kilocycles. The tank circuit consists of L1 in parallel with the three series capacitors, C1, C2, and C3. Oscillations are sustained by generative feedback voltages obtained at the junction of C1 and C2 and the junction of C2 and C3. Cathode resistor R2 provides protective bias for both sections of V1. The feedback voltages are developed across grid resistors R1 and R3. The center tap of L1 is returned to a voltage divider between plus 250 volts and ground. Adjustment of R5 varies the amplitude of the output signal. R5 is adjusted to obtain 70 volts rms at TP1. C4 provides an a-c ground at the centertap of L1.

c. Power amplifiers V2 and V3. This stage is a push-pull amplifier which also serves to isolate the oscillator from the load. This prevents any changes in the load from affecting the frequency of the output signal. The outputs of V1 are applied through the input resistors, R7 and R12. Capacitor C5 and resistor R9 form a coupling circuit for V2; C6 and R11 form the coupling circuit for V3. Common cathode resistors R10 and R19 in parallel establish the proper bias. Negative feedback is obtained through resistors R8 and R13, connected between the plate and grid circuits. The relative size of these feedback resistors and the input resistors, R7 and R12, determines the gain of the power amplifier. The voltage gain is equal to the ratio of feedback resistance to input resistance, and hence is equal to 1.2 in this circuit. However, power gain is considerably greater. The push-pull signal is applied to the primary of transformer T1, which is centertapped to plus 250 volts. Two of the three

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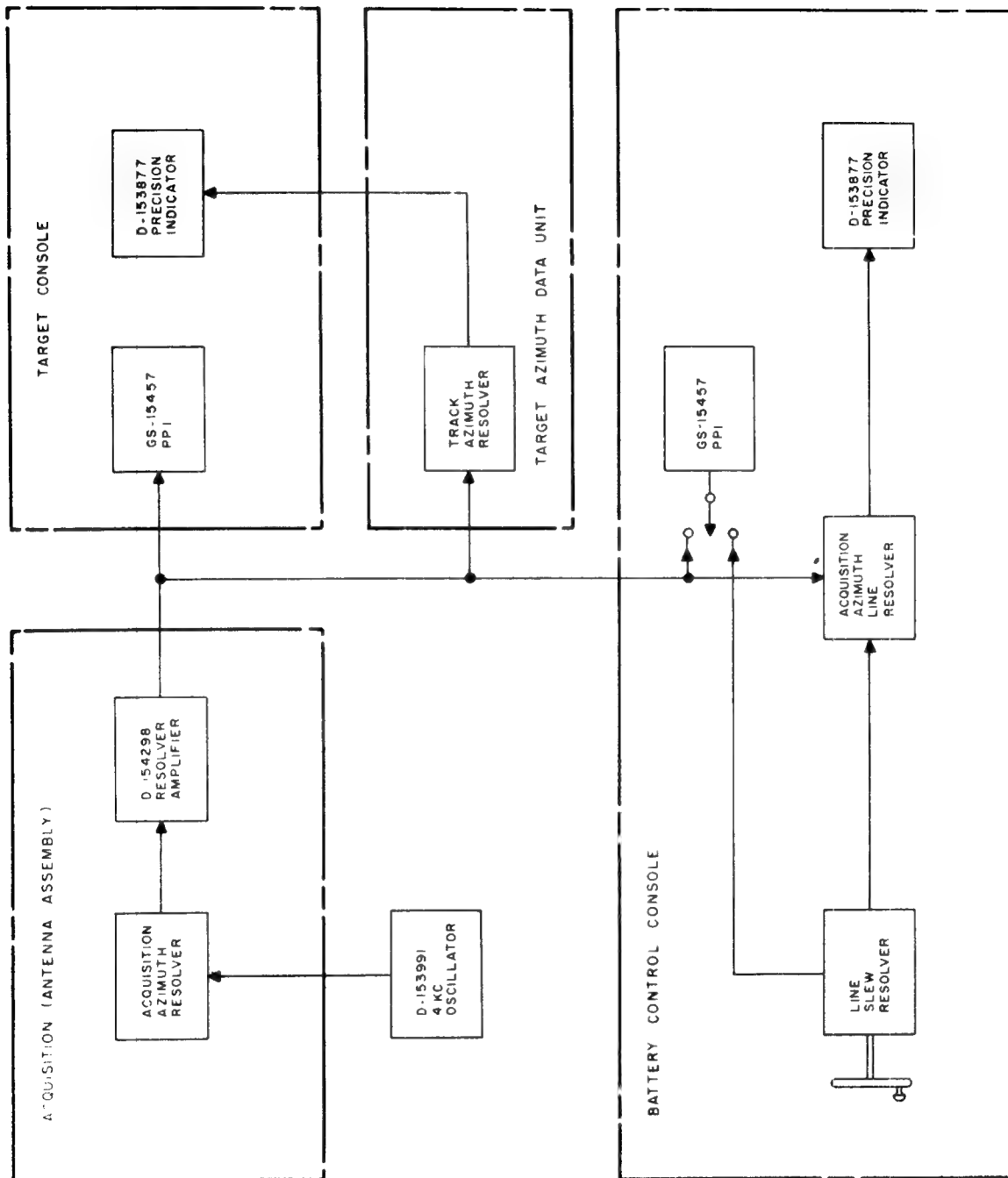


Figure 29. Acquisition resolvers.

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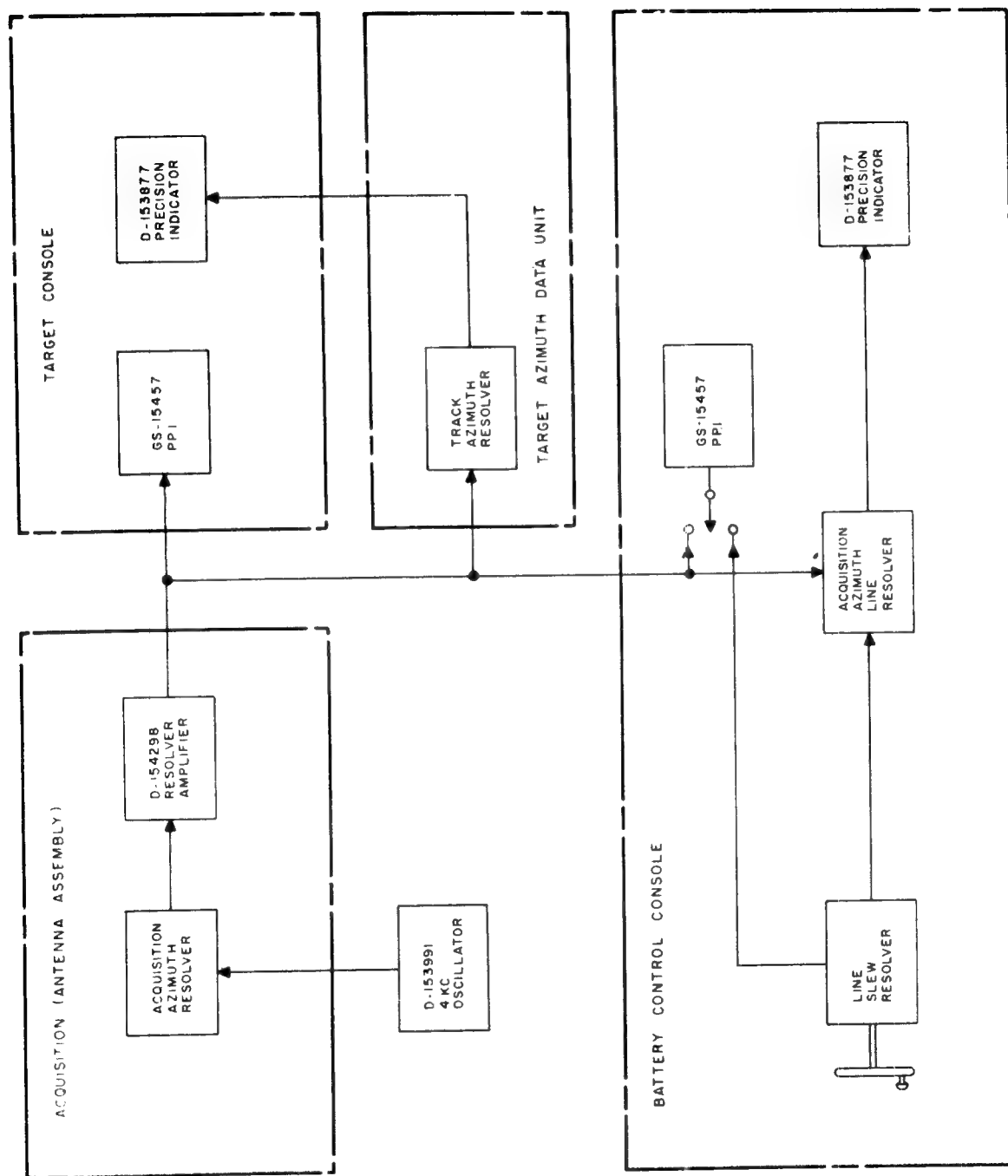


Figure 29. Acquisition resolvers.

91. AZIMUTH CHANNEL

a. Acquisition azimuth resolver, general (TM 9-5000-25, sheet 25). The purpose of the acquisition azimuth resolver B2 is to transmit the angular position of the antenna to the trailer (fig 29). One of the 4-kc outputs of the 4-kc oscillator, designated ACQ 4-kc is applied to one of the stators of the acquisition azimuth resolver whose rotor is mechanically coupled to and rotates in synchronism with the acquisition antenna. The waveform appearing across the N-S winding of the acquisition azimuth resolver rotor coil is a 4-kc signal with an amplitude that varies as the sine of the acquisition antenna azimuth angle. These two position signals are applied to the resolver amplifier chassis for impedance matching.

b. Acquisition azimuth resolver operation (TM 9-5000-25, sheet 25). The acquisition azimuth resolver stator is excited by the 4-kc acquisition signal, which is 25 volts rms. The stators of this resolver are connected in such a manner that S1-S3 is excited and S2-S4 is shorted out. The output of rotor R1 will be a 4-kc signal varying in amplitude between 0 and 25 volts and being maximum when the antenna is pointing north or south. The output on R2 is the same type of signal, except that its maximums occur when the antenna is pointing east or west. These two outputs are always 90° out of phase. The output voltage from this resolver would be loaded down if sent to the trailer circuits directly, so it is first amplified in the resolver amplifier chassis.

c. Resolver amplifier chassis, general (TM 9-5000-26, page 224). The resolver amplifier chassis is located in the acquisition antenna drive unit. Its function is to match the impedance of the acquisition azimuth resolver to the impedance of the cable and subsequent circuits and components and thus isolate the acquisition azimuth resolver from those components. The two signals from the acquisition azimuth resolver provide the inputs. These voltages pass through two identical channels. The gain of these channels is slightly greater than unity. Both of the applied signals experience a phase reversal of 180° in the channels. The output signals are proportional to the N-S and E-W components of the acquisition azimuth angle. The outputs of these channels are connected to the tracking azimuth resolver, the azimuth line resolver, and the ring demodulators in the acquisition radar presentation system. Each channel of the resolver amplifier consists of a voltage amplifier and a power amplifier. Negative feedback limits the gain to a low value. Transformer coupling is employed in the output. The plate voltage for the resolver amplifier chassis is connected through the acquisition azimuth scan switches so that in order to get an output from this chassis the switches must be turned on. Since stages V3 and V4 are identical to V1 and V2, only V1 and V2 will be discussed.

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d. Voltage amplifier V1. The N-S output voltage of the acquisition azimuth resolver is applied at one end of a signal voltage divider consisting of R1 and R2 and C6 and C7. The reactance of the capacitors at the frequency of 4 kilocycles is so high that only the resistance of R1 and R2 need be considered in determining the signal amplitude appearing at the grid. The capacitors prevent distortion of the signal by causing the time constants of each section of the voltage divider to be equal. A signal opposite in phase to the input signal is applied at the R2 end of the network. The signal which appears at the control grid of V1 is in phase with the N-S voltage output of the acquisition azimuth resolver, but is considerably reduced in amplitude. The cathode bypass capacitor, C5, is considerably smaller than is usual in a voltage amplifier stage. At the operating frequency, C5 establishes the cathode. However, at frequencies below 4 kilocycles, the reactance of C5 increases. The resulting degeneration attenuates those lower frequencies. In other respects, V1 is a conventional amplifier.

e. Power amplifier V2. This stage is designed to permit heavy current flow in the primary of transformer T2, which forms the plate load. Resistor R7 prevents parasitic oscillations. The cathode circuit is similar to the cathode circuit of the preceding stage. The connections made to the plate transformer are such that the output signal is opposite in phase to the signal appearing at the plate of V2. The circuit consisting of C9 and R10 across the primary and C4 and R21 across the secondary of the transformer is designed to tune the transformer to 4 kilocycles. If the input signal from the acquisition azimuth resolver at a certain instant is positive, then a negative signal will appear at the plate of V1 and at the grid of V2, a positive signal will appear at the plate of V2, and a negative signal will appear at the output. This negative signal, 180° out of phase with the input signal, is applied as feedback to the R2 end of the frequency-compensated voltage divider. One effect of this feedback is the reduction of the over-all gain of the two stages V1 and V2 to a value slightly greater than unity. As a result, high fidelity reproduction of the signal is obtained. A second effect is the reduction of the output impedance to an apparent value of slightly over one ohm. This feature insures that the signals produced in the acquisition azimuth resolver will be unaffected by any changes in the load.

f. Output distribution. The outputs of the azimuth channel are distributed to four points after the signals reach the trailer. The track azimuth resolver in the azimuth data converter receives these signals, and they are used to determine the azimuth at which the electronic cross appears. The azimuth line resolvers receive this signal also, and the signal is used to determine the azimuth of the flashing steerable azimuth line. The ring demodulators in the acquisition presentation system also receive these signals to form the rotating sweep. Before the signal reaches the PPI's, however, it must pass through the contacts of relay K(7D)1 and K(7E)1 panel.

Section III. ACQUISITION RANGE UNIT

92. INTRODUCTION (TM 9-5000-26, page 225 and TM 9-5000-25, sheets 24 and 32)

The purpose of the acquisition range control is to allow the range circle to be positioned from the target designate control panel. The range of the range circle is determined by the voltage tapped off potentiometer R1 in the acquisition control panel (TM 9-5000-26, page 223). This potentiometer and its range counter may be turned directly by the acquisition operator or by the battery control officer. To perform this remote control job, a servo system called the acquisition range control is used. This servo system is arranged in such a manner that the control point is easily accessible to either the acquisition operator or the battery control officer.

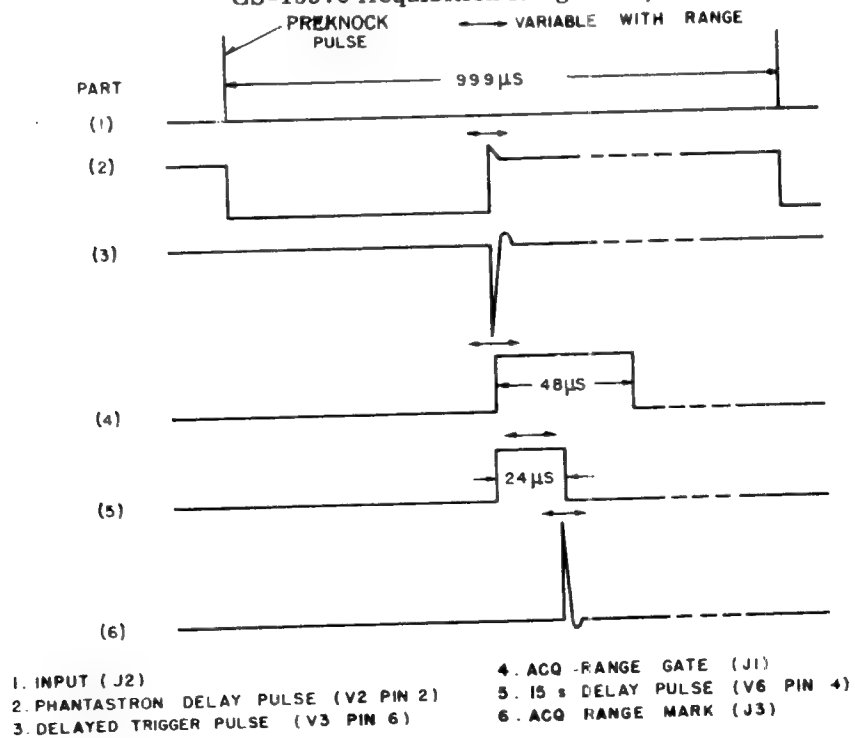
93. GENERAL

a. Range detection. The range of a target detected by the acquisition radar is determined by measuring the time interval between the transmission of the pulse (main bang) and reception of the echo from the target in question. Extreme accuracy is unnecessary for this time measurement, since acquisition range data are used only for target designation and other surveillance functions. The accuracy requirement for the acquisition range determination system is ± 150 yards, corresponding to a time measurement accuracy of ± 0.9 microseconds.

b. Method of time measurement. The time measurement is made by aligning a continuously movable range marker, the range circle, with the target pip. This range mark is generated in the acquisition range unit. It is a sharp pulse occurring after each main bang. The time delay between the main bang and the range mark is controlled by a potentiometer geared to a dial, calibrated in yards. The range mark is displayed on the B-indicator, in the battery control console (fig 2). On the B-indicator the range mark appears as a horizontal line.

c. Range potentiometer. The range potentiometer is installed in the target designate control panel, which is located in the battery control drawer. The potentiometer is geared directly to the handwheel on the control drawer marked RANGE and to the acquisition range dial on the control drawer.

d. Range unit (fig 30). The acquisition range unit is connected electrically to the range potentiometer in the acquisition range control unit. The circuits of the acquisition range unit are triggered by the preknock pulse, generated by the synchronizer 23.5 microseconds before the main bang. The unit has



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two outputs (fig 30); the acquisition range mark, a sharp positive pulse whose delay with respect to the input trigger is controlled by the range potentiometer and may vary between about 44 microseconds and 756 microseconds, corresponding to ranges between 3,500 and 120,000 yards; and the acquisition range gate, an 8,000-yard, positive pulse whose leading slope precedes the acquisition range mark by 28.5 microseconds.

e. Output distribution (TM 9-5000-25, sheet 24). The acquisition range mark is supplied to the PPI's in the target console and in the battery control console, where it is displayed as the range circle, and to the video and mark mixer in the acquisition radar assembly, where it is used in the precision indicator on the battery control console for display as the acquisition range line. It is fed to the target range slew control unit where it is compared electronically, in respect to time of occurrence, with the acquisition-track range mark from the tracking range unit. This comparison of relative time occurrence informs the slew control unit whether to slew in or out in range to find the designated target. The detailed function of this slew control unit is shown in TM 9-5000-26. The acquisition range gate is supplied to the precision indicator in the battery control console where it is used for generating the expanded range sweep on the B-scan display.

f. Acquisition range gate channel. Both of the inputs to the acquisition range computer are applied to this channel. A phantastron stage, triggered by the preknock pulse, produces a variable-width signal. The trailing edge of this signal triggers a multivibrator which generates the acquisition range gate. This gate signal is 30 microseconds wide, corresponding to a range of 8,000 yards. It may be delayed linearly in range from a minimum of 3,500 yards to a maximum of approximately 120,000 yards. This signal is applied to the battery control precision indicator where it is used to produce the range sweeps and to aid in unblanking the scopes.

g. Acquisition range mark channel. The input to this channel is the acquisition range gate. From this signal, the channel produces a positive 0.5-microsecond pulse that has a fixed delay of 15 microseconds from the starting time of the gate. As the acquisition range gate is moved in range, the acquisition range mark also moves, always occurring 15 microseconds or 2,500 yards after the leading edge of the gate. When the acquisition range gate is adjusted to the proper width, the acquisition range mark occurs at the center of the gate. The limits of movement of the acquisition range mark are from 3,500 yards to 120,000 yards. This signal is applied to the mixer channel and to the relay amplifier channel. It produces the range circle, which is used to designate the range of targets appearing on the PPI's. In the target range slew control unit, the signal is used for comparison with the acquisition track range mark. This comparison is used to position the target track range system to

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the range of designated targets. When the outer ring of the AZIMUTH LINE control is depressed, the acquisition range mark is applied directly to the PPI's in place of mixed video and marks.

94. PHANTASTRON

a. General. The phantastron stage consists of V1, V2, and V3A. The preknock pulse is applied through V1A, which acts as an isolation stage between the pulse synchronizer and the phantastron. Phantastron V2 is similar to a one-shot, cathode-coupled multivibrator. It is triggered by the preknock pulse from V1A. The output of V2 is a negative square wave whose positive-going trailing edge can be made to vary linearly in time by a change in the d-c control voltage applied through V1B. V1B is an isolation stage whose input is the d-c control voltage from the acquisition range potentiometer, R6. The range potentiometer is connected through a servo loop to the two RANGE CIRCLE handwheels. V3A is used to insure quick recovery of V2 between preknock pulses.

- (1) Diode V1B determines the quiescent plate voltage of the phantastron in accordance with the amplitude of the d-c control voltage obtained from R6. The cathode is connected to RANGE potentiometer R6, and the plate is connected to the plate of the phantastron. The RANGE potentiometer is connected to plus 150 volts and is grounded through ZERO control R2 and resistor R3. R2 establishes the minimum d-c level which may be obtained from the brush arm of R6. Capacitor C18B is a bypass capacitor which operates to maintain a constant level in the d-c output of R6.
- (2) Diode V1A is an isolating diode through which the triggering signal, the preknock pulse, is applied to the phantastron. The preknock pulse is developed across R1 and is applied to the plate of the diode. The cathode is held at a d-c level of plus 4.4 volts by voltage divider, R4-R5, and is normally cut off. The positive preknock pulse allows the diode to conduct, and the signal appears at the cathode of V1A and at the suppressor grid of the phantastron.
- (3) Phantastron V2 consists of a large 1-megohm plate-load resistor which maintains the linear response of the stage. The plate is connected to the grid of triode V3A. The screen grid is connected to a positive potential of plus 86 volts at voltage divider R6-R10. The control grid is returned to plus 150 volts through resistors R12 and R14 in series with R13, the rate potentiometer. Bypass capacitor C1A keeps fluctuations in grid voltage out of the power supply. The output

of the stage is taken across cathode resistor R11 and is applied through the R-C coupling circuit consisting of C2 and R19 to the grid of differentiator-amplifier V3B.

- (4) Triode V3A provides a fast-charge path for C3 during recovery of the phantastron. The voltage present at the phantastron plate is applied to the control grid of the triode, and an almost equal voltage is developed across cathode resistor R15. When the d-c control voltage at R6 is 150 volts, the potential at the cathode of the triode will be nearly the same.

b. Quiescent period. The state of the phantastron stage during the quiescent period is as follows: V1B is conducting at a level determined by the setting of range potentiometer R6. The plate of V2 is cut off by the suppressor grid, and V1B will set the voltage level of the plate of V2. V1A is cut off by the positive d-c potential at the cathode. The plate of V2 is cut off by the low d-c potential at the suppressor grid and the high potential at the cathode, which is due to the heavy conduction of the screen grid and control grid. V3A is conducting at a rate determined by the plate voltage of V2, which is connected to the control grid. At the cathode this level of conduction develops a corresponding voltage which is applied to C3. Thus the charge across C3 is the difference between the d-c level of the control grid of V2 and the d-c level of the cathode of V3A. No plate current is flowing in the phantastron, but screen-grid and control-grid current is considerable.

c. Signal period. The preknock pulse is applied to the plate of V1A causing it to conduct. A positive signal is developed at the cathode of the diode and is applied to the suppressor grid of V2. This positive signal overcomes the bias between the suppressor grid and the cathode of V2, and current begins to flow to the plate (at the expense of the screen-grid current). This causes the plate voltage to drop. This drop in voltage is coupled directly to the grid of V3A, causing that tube to conduct less and reducing the voltage across cathode resistor R15. This drop is coupled by C3 to the control grid of V2. The drop in voltage at the control grid causes V2 to conduct less. With V2 conducting less, there is less current flow in the cathode circuit and less voltage is developed across cathode resistor R11. With less potential on the cathode, the bias between the suppressor grid and the cathode is less, and the flow of current to the plate (at the expense of the screen-grid and the control-grid current) is increased. The voltage drop at the plate of V2 increases, and the cycle continues cumulatively and almost instantaneously until any further drop in potential on the control grid of V2 will tend to cut the tube off altogether. At this instant, C3 starts to discharge. The discharge path is through R12, R13, C1A, ground, and R15 to the other side of C3. However, the rate of discharge is controlled by the linear rate of current flow through plate-load resistor R9, which controls the conduction of V3A. V2 will conduct in this lower state, through the plate,

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until the control grid rises (as C3 discharges linearly) to a point where the cathode is at such a potential relative to the suppressor grid that a high bias relationship is established. At that time, a cumulative and almost instantaneous action takes place. Screen-grid current starts to increase, further increasing the potential at the cathode, which in turn further increases the bias between cathode and suppressor grid. This allows less current to flow to the plate. The plate voltage rises and this rise is coupled to the grid of V3A, whose cathode rises correspondingly. This immediate rise is coupled by C3 to the control grid whose rise will further increase the conduction of V2 and thereby increase the potential at the cathode. The cycle continues until the control grid begins to draw current and the plate of V2 is cut off. C3 must then charge to the difference in d-c voltage between the control grid and the plate (less the small drop encountered across V3A). The charge path of C3 is through V3A, to ground through the power supply, the cathode of V2 to the control grid, to the other side of C3. The output during the signal stage is a negative-going waveform whose trailing edge may be varied linearly in time by a change in the d-c level at the plate of V2. By increasing the d-c level at the plate of V2 the charge across C3 is increased correspondingly. Then, upon application of the preknock pulse, the drop at the plate will be greater because of the higher initial voltage level. Thus the eventual drop coupled across C3 will be greater, and the capacitor will require a longer time of discharge to reach the point where the tube will start its cumulative action and return to the quiescent state. The reverse effect will take place with a decrease in the d-c level at the plate of V2.

95. DIFFERENTIATOR-AMPLIFIER

The differentiator-amplifier stage consisting of V3B and CR1 amplifies, inverts, and differentiates the positive-going trailing edge of the square-wave output of the phantastron stage. The plate of this stage is connected to the positive 150-volt supply through plate-load resistor R18. The grid of V3B is connected to the plate through an R-C circuit which provides negative feedback. The input signal to the grid is the negative-going waveform from the cathode of V2, applied through the coupling circuit consisting of C2 and R19. The cathode of V3B produces self-bias through the action of R16 and C6. With the grid returned to ground, there is sufficient voltage developed across R16 to keep the tube barely above cutoff. Application of the negative square wave to the control grid cuts the tube off. This causes a small positive rise at the plate. This positive signal is blocked from the next stage by CR1, but is immediately coupled back through C5 and R17 to the control grid. This rise opposes and almost cancels the drop previously present on the grid. The action continues for a very short period of time, and the tube returns quickly to its previous level near cutoff. The trailing edge of the negative square wave from V2 is positive-going. When it arrives at the grid of V3B, the tube will

conduct more heavily. This will cause a drop at the plate which is coupled back to the grid through C5 and R17, causing the tube to be cut off. After a series of quick approximations, the tube resumes its previous level near cut-off. At the plate, the sharp negative drop and resultant rise (due to feedback) form a narrow negative pulse which is coupled through C4. The polarization of CR1 is such that this negative signal is allowed to pass through the grid of V4A.

96. GATE MULTIVIBRATOR V4

The gate multivibrator stage is a cathode-coupled, one-shot multivibrator which is triggered by the negative-spike output of the differentiator-amplifier state. The output of V4 is a positive square wave whose width is adjusted to 5,000 yards, or 30 microseconds. The plate, pin 6, is connected to the plus 250-volt supply through plate-load resistor R24. C1B and R23 form a decoupling network to bypass voltage fluctuations to ground. C7, R21, and R20 form a coupling circuit from the plate to the grid, pin 2. The plate, pin 1, is connected to the plus 250-volt supply through plate-load resistor R22. The grid, pin 2, is connected to the same supply through R21 and R20. These resistors, together with parallel capacitor C7, control the discharge time constant which determines the width of the square wave output. The cathodes are connected to ground through a parallel resistor network, consisting of R25, R26, R27, and R28, which together total approximately 3,000 ohms. The grid, pin 7, is connected so that it is at a potential approximately equal to two-thirds the potential on the cathodes when the A-section is conducting. C9 keeps this potential fairly constant and bypasses voltage fluctuations to ground. R29 is a grid-limiting resistor. During the quiescent state of V4, the A-section is conducting, and the B-section is cut off. The grid, pin 2, is drawing current. The high level of current flow in the A-section develops a comparatively large voltage at the cathode. The bias between cathode and grid of the B-section is always one-third the potential at the cathode. When the A-section is conducting, this keeps the B-section cut off. C7 is charged to the difference of potential between the grid of the A-section, which is drawing current, and the plate of the B-section, which is at the supply potential. When the negative peak from the differentiator-amplifier is impressed on the grid of the A-section, the A-section conducts less. The voltage at the cathode decreases because of the reduced current flow through the tube. The potential at the grid of the B-section remains relatively constant, however, because of the decoupling effect of C9 and R28. This drop in the cathode potential relative to the constant potential at the grid of the B-section overcomes the cutoff bias of the B-section and allows it to conduct. The voltage at the plate of the B-section drops, and this drop is coupled to the grid of the A-section through C7, causing the A-section to be cut off. Capacitor C7 then begins to discharge through R21 and R20. When C7 discharges to the cutoff level of the A-section, that section will begin to conduct. This will cause

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an increase of current flow through the cathode circuit, raising the cathode potential. When the cathode potential rises to the point where the one-third differential between cathode and grid is greater than the cutoff potential of the tube, the B-section will be cut off. C7 will then charge through R24, C1B, ground, from ground through the cathode, to the grid of the A-section (which is now drawing current), to the other side of C7. A positive square-wave signal is taken from the plate of the A-section and is coupled through C10 to V5B. The same signal is coupled through C12 to V5A. The width of the square wave is adjusted to 30 microseconds by GATE LENGTH control R20. This control varies the discharge time of C7.

97. CATHODE FOLLOWER V5B

Cathode follower V5B is the output stage of the acquisition range gate channel. It is used for impedance matching. The tube is biased at cutoff, and grid limiting takes place. Thus, the stage improves the waveshape of the positive square-wave output of the acquisition range gate channel. This output is applied to the des-ignator control system to generate the range sweeps and for use as an unblanking voltage when the TRACK-ACQ switch is in the ACQ position. The plate is connected to the supply through the decoupling network, R 38 and C18C, which maintains a constant d-c level at the plate. The cathode is connected to ground through resistor R34. The output of the acquisition range gate channel is developed across R34 and applied to J1. The grid is connected to minus 24 volts through voltage divider R31-R32. C11 is a bypass capacitor which maintains the d-c level at the junction of R31 and R32 at a constant value. R33 is a grid-limiting resistor. With minus 24 volts on the grid and the cathode at ground, V5B is cut off. The positive square wave from V4 is coupled through C10 and across R30 and is impressed on the grid. This signal overcomes the cutoff bias, and the tube conducts. The amplitude of the positive square wave applied to the grid is sufficient to cause grid current to flow, resulting in grid limiting.

98. L-C DELAY NETWORK

The L-C delay stage consists of V5A, Z1, and V6A. The input is the positive square wave from gate multivibrator V4 in the acquisition range gate channel. V5A is a driver for the L-C circuit of Z1. Prior to a signal, V5A is cut off. With the positive square wave applied, V5A conducts, inverting the signal and causing Z1 to oscillate at its natural frequency of 16.67 kilocycles. Z1 is a series-resonant L-C circuit which is damped by the action of V6A. V6A allows Z1 to oscillate only for one-quarter of a cycle. This stage is factory-adjusted to develop a signal exactly 15 microseconds in width. Thus the output of this stage is a positive waveform 15 microseconds wide. The cathode of driver tube V5A is connected to ground. The plate is connected to the plus 150-volt supply through plate-load resistor R39. The plate is also connected to L-C

circuit Z1. The grid is connected through R35 to voltage-divider network R37 and R36, placing the grid at a potential of -40 volts. C13 is a bypass capacitor which maintains a constant d-c level at the junction of R36 and R37. Z1 is an L-C series circuit, resonant at 16.67 kilocycles. The plates of the capacitors are at +150 volts on one side and are connected to ground through the coil on the other. This places a charge of 150 volts across the capacitors during the no-signal period. The cathode of V6A is connected to ground. The plate is connected to the +15-volt supply through plate-load resistor R40. R44 and C18A form a decoupling network to maintain the d-c voltage at a constant level. The grid of V6A is connected to the junction of the capacitors and the coil of Z1. During the quiescent state, V5A is cut off. This places +150 volts on the plate and on one side of the capacitors of Z1. V6A is conducting heavily. The grid is not drawing current at this time being at ground potential. With the positive square wave from the multivibrator applied to the grid of V5A, that tube conducts, and the voltage at the plate drops. This drop is reflected across the capacitors of Z1 to the grid of V6A, which is driven below ground. The capacitors start to discharge through the coil to ground, completing the discharge circuit through V5A. Z1 discharges at resonance, and 15 microseconds after the initial signal the capacitors will have discharged to ground potential through the coil. In discharging through the coil, the current builds up a magnetic field about the coil. This field now begins to collapse, causing current flow to continue in the same direction. This drives the grid of V6A above ground, causing grid current to flow. This grid current damps out further oscillations. The appearance of the negative-going trailing edge of the square wave from V4 at the grid of V5A will restore the tube cutoff. The plate will then return to the supply potential, and the capacitors of Z1 will charge through R39 to the supply voltage. During the 15-microsecond negative pulse at the grid of V6A, the conduction through the tube will be reduced.

99. PULSE-SHARING AMPLIFIER

Cutoff limiting is used to square off the output at the plate of V6A. Therefore, the output taken from the plate will be a positive square wave 15 microseconds in duration. The output is coupled through C14 to V6B. V6B is a pulse-shaping amplifier biased below cutoff. The cathode is at ground. The grid is connected to -44 volts through voltage-divider network R41 and R42 in order to prevent any spurious pips at the trailing edge of the input to V6B from appearing at its plate. The positive waveform from V6A is applied to the grid and overcomes the cutoff bias. The plate voltage drops, and the output signal is a negative square wave 15 microseconds wide. This output is differentiated by C16 and R46 to produce a negative peak corresponding to the negative-going leading edge of the square wave, and a positive peak, 15 microseconds later, corresponding to the positive-going trailing edge. These negative and positive peaks are applied to the control grid of amplifier V7.

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100. AMPLIFIER V7

The final stage in the acquisition range mark channel is amplifier V7 which has an R-C peaking circuit in the input and utilizes a transformer plate load. The tube is almost cut off prior to the signal from V6B. Being almost cut off, the tube will conduct only when the peaked, positive-going, trailing edge appears. The output is coupled through transformer T1. From the secondary of T1, a positive 0.5-microsecond spike is obtained. This is the acquisition range mark. This output is applied to the relay amplifier channel, acquisition presentation system, the trial fire indicator channel, and the mixer channel. The plate is connected to +250 volts through R47, which prevents oscillations caused by T1 and C1C. T1 inverts the signal and couples it to J3. C1C and R47 form a decoupling network to maintain a constant d-c level at the screen grid. The screen is at approximately the same potential as the plate. The cathode is at the same potential as the suppressor grid. C17 maintains a constant d-c level at the cathode and the suppressor grid. The control grid is at ground potential, thus maintaining a high bias relationship to the cathode. The negative-going peak applied to the control grid of V7 has little effect on the conduction of the tube, as it is almost cut off. The positive-going peak, however, causes the tube to conduct. The voltage at the plate drops. This negative pulse is coupled through transformer T1, appearing as a positive 0.5-microsecond pulse across the secondary. The output signal leaves the unit at J3.

Section IV. SWEEP CHANNELS HIGH-VOLTAGE POWER SUPPLIES

101. GENERAL (TM 9-5000-26, pages 227, 228)

a. Ring demodulators. The purpose of the ring demodulators is to remove the envelope from the azimuth channel signal so that the amplitude of this signal can be used to control sweep direction. For each PPI there is a N-S ring demodulator. The output voltages of the azimuth channel to the ring demodulators, and the 4-kc carrier is also applied as an input. The amplitude of the N-S signal indicates the magnitude of the N-S component of the acquisition azimuth angle, being greatest when that angle is either due north or due south. The direction of the N-S component of the acquisition angle is determined by the phase relationship between the N-S resolver signal and the 4-kc carrier, being in a northerly direction when an inphase relationship exists and in a southerly direction when an out-of-phase relationship exists. The N-S ring demodulator detects the modulation and phase of the applied N-S resolver signal and produces 2 output sine waves, a north voltage and a south voltage. The 2 sine waves are 180° out of phase. The north voltage attains maximum amplitude when the acquisition azimuth is due north and the south voltage attains maximum amplitude when the azimuth is due south. The E-W ring demodulator functions

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in a similar manner to produce an east voltage and a west voltage. The demodulator chassis consists primarily of a ring demodulator, 2 R-C filter circuits, and 2 output cathode followers. Each of these units detects the modulation of the resolver signal applied as an input and produces as an output 2 d-c voltages which vary sinusoidally in accordance with this modulation. The sinusoidal variation of these voltages is in phase opposition. Both ride at a positive d-c level so that the sinusoidal variation is within the approximate limits of +3 volts and +30 volts. The 4-kc component of the signal is removed in the R-C filter circuits. The two cathode followers isolate the ring demodulator from the circuits to which the output signal is applied. The N-S ring demodulator receives the N-S azimuth channel voltage, and its two outputs are a north voltage and a south voltage. These signals, together with an east voltage and a west voltage produced by the E-W ring demodulator, are applied to the sweep generator.

b. Range gate channel (TM 9-5000-26, page 231). The input to the range gate channel is the preknock pulse. This channel uses the preknock pulse to trigger a multivibrator whose outputs are 2 positive square waves. One of these outputs is used to unblank the PPI tube; the other output is used in the sweep generators to aid in the generation of the sweep voltages.

c. Sweep generators (TM 9-5000-26, page 230). The sweep gate of 60,000 yards, or 120,000 yards, is applied to the 4 identical generators. These 4 stages generate sweep voltages during the application of the sweep gate signal. The north, south, east, and west voltages from the demodulators are also applied to the sweep generators. These 4 inputs determine the amplitude of the 4 sweep voltages. As a result, the amplitude of the north sweep voltage is maximum when the acquisition azimuth is due north and is minimum when the azimuth is due south. The east sweep voltage is of maximum amplitude when the acquisition antenna is at an azimuth of 1,600 mils (due east) and is minimum when the azimuth is due west. Similarly, the south sweep voltage and west sweep voltage developed in the sweep generator vary in amplitude as the acquisition antenna rotates. The amplitude of the north sweep voltage varies as the cosine of the azimuth angle, and the amplitude of the east sweep voltage varies as the sine of the azimuth angle. The modulation of the south and west sweep voltage is 180° out of phase with that of the north and east sweep voltage, respectively.

d. Sweep channel amplifiers (TM 9-5000-26, page 232). Each sweep channel contains 6 amplifier stages, 4 voltage amplifiers, and 2 power amplifiers. The output of the north sweep generator V2A is applied to voltage amplifiers V5 and V6 where it receives sufficient voltage amplification to operate the power amplifier V7. Power amplifier V7 amplifies its input to a sufficient power level to drive the deflection coil. The south sweep generator has its

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output amplifier in a similar manner in stages V1, V2, and V4. The east-west sweep channel operates in an identical manner to the north-south sweep channel. The north-south coils are so positioned that the magnetic field generated about them will deflect the PPI sweep vertically. The amount of vertical deflection is determined by the relative amplitude of north and south sweep voltages, and the direction of vertical deflection is determined by the sweep voltage having greater amplitude. In a similar fashion, horizontal deflection of the PPI sweep is determined both in direction and in extent by the east and west sweep voltage outputs of the E-W sweep amplifier, which are applied to a second pair of opposing deflection coils. The direction in which the sweep is deflected is the vectorial result of the forces which may be considered as acting only in horizontal and vertical directions (although an oversimplification, such a concept is useful for purposes of explanation). As the antenna rotates, the four sweep voltages vary in amplitude in such a manner as to cause the PPI sweep to be deflected in the direction which corresponds to the azimuth of the antenna, and the sweep is made to rotate in synchronism with the antenna.

e. PPI video amplifier. The PPI video amplifier consists of one stage of video amplification. The input to this stage is normally the video and marks from the video and mark channel. The output negative signals are applied to the cathode of the PPI tube.

f. PPI tube. The PPI tube is a 10-inch 10KP7 cathode ray tube. Deflection of the beam is accomplished by the flow of sweep currents through stationary magnetic deflection coils. A permanent magnet is employed for focusing and centering. The high voltage required for operation of the two PPI tubes is produced by a voltage-doubler power supply. The required potential is 8,500 volts. The tube is shown schematically on page 231 of TM 9-5000-26

g. Track and acquisition indicator HV power supplies. The indicator HV supply furnishes the HV used on all cathode ray tubes of the acquisition and track radars. The supply produces +8,500 volts for the PPI's, 5,000 volts for the PI's, and +2,000 volts direct current, for use in the tracking scopes and the PPI.

102. SWEEP CHANNEL, RING DEMODULATORS (TM 9-5000-26, page 229).

a. Ring demodulator, basic considerations. The 4-kc carrier is applied to the primary of transformer T2. The N-S or the E-W azimuth channel signal, a modulated 4-kc sine wave, is applied through AMPLITUDE control R1 to the primary of transformer T1. The 2 transformers are similar, but the connections are such that the step-up ratio of T2 is double that of T1. T2 has an effective step-up ratio of 1:5 and T1 has 1:2.5. The voltage induced in

the secondaries of both transformers are applied across the bridge circuit consisting of diodes V1A, V1B, V2A, and V2B, and their cathode resistors. Because of the difference in the turns ratios of the two transformers, the reference voltage (the 4-kc carrier) is of much greater amplitude. This condition is necessary for the proper operation of the circuit. At an instant when the phase of the carrier is such that pin 4 of T2 is positive with respect to pin 6 current will flow from pin 6 of T2 through V2A and V1B to pin 4 of T2. During the next half-cycle of the 4-kc signal across the secondary of T2, current will flow from pin 4 through diode V2B and V1A to pin 6. The key to the operation of the ring demodulator, however, is in the current flow through the secondary of transformer T1. Current can flow in only half of the secondary of T1 at any one instant. In this respect, T1 is similar to the plate transformer of a conventional full-wave rectifier in which the current path is completed through the centertap connection of the secondary. At an instant when current enters the secondary of T1 at either pin 4 or pin 6, it will pass to ground through R6, developing a negative signal across R6. The circuit is completed through R7 at the centertap of T2, and the current flow through R7 will develop a positive signal across that resistor. At an instant when current leaves T1 at either pin 4 or pin 6, it must enter through R6 at the centertap. In this condition, the circuit is again completed through R7 in the secondary of T2. However, the direction of current flow through R6 and R7 is now opposite to what it was in the first example, and opposite phase voltages are developed across these resistors.

b. Inphase operation, positive half-cycle. At the instant when the resolver signal and the carrier are in phase, inphase signals will appear at pin 4 of T1 and pin 4 of T2. When pin 4 of T1 is positive with respect to pin 6 of T1, and pin 4 of T2 is positive with respect to pin 6 of T2, current will flow from pin 6 of T2 through diodes V2A and V1B to pin 4 of T2. At transformer T1, pin 6 is the most negative terminal. However, current is prevented from flowing out of the secondary through this terminal. Current cannot pass through diode V2B in the reverse direction. Current cannot pass through diode V1A because the plate of V1A is connected to terminal 6 of T2, and that point is more negative than terminal 6 of T1. This is true because the amplitude of the carrier signal across the secondary of T2 is greater than the amplitude of the resolver signal across the secondary of T1. The current flow which results from the voltage across the secondary of T1 must therefore follow the path from pin 5 of T1 to ground through R6 (developing a negative signal), from ground through R7 (developing a positive signal), to pin 5 of T2, to pin 6 of T2, and through diode V2A to pin 4 of T1.

c. Inphase operation, negative half-cycle. During the next half-cycle of the applied 4-kc signals, pin 4 of T1 becomes negative with respect to pin 5 of T1, and pin 4 of T2 becomes negative with respect to pin 6 of T2. However, because the two signals applied to the circuit are still in phase, there should be no change in the polarity of the voltage developed across resistors R6 and R7.

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As a result of the reference voltage induced in the secondary of T2, current will flow from pin 4 through diodes V2B and V1A to pin 6 of T2. Although the most negative terminal of T1 is now pin 4, current cannot flow from pin 4 of T1 because the plate of V1B is connected to pin 4 of T2 and is therefore more negative than the cathode. Current therefore leaves T1 at the centertap connection, passes to ground through R6 (again developing a negative signal), from ground through R7 (again developing a positive signal), to pin 5 of T2, and from pin 4 of T2 through diode V2B to pin 6 of T1. The current flow through R6 and R7 is determined by the voltages applied to the circuit. As the two 4-kc signals pass through zero value, each half-cycle of the 4-kc voltage, current flow through R6 and R7 also falls to zero. The voltage developed across each resistor, therefore, is a pulsating d-c voltage such as would be obtained from a conventional full-wave rectifier. The amount of current flow through R6 and R7 is determined by the amplitude of the modulated 4-kc resolver signal applied to transformer T1, and hence the amplitude of the pulsating d-c voltage developed across the resistors is proportional to the amplitude of the N-S resolver signal. While the N-S resolver signal is in phase with the 4-kc carrier, the acquisition antenna makes a half revolution, and during this period the resolver signal passes from zero amplitude through maximum amplitude and back to zero amplitude. Therefore, the pulsating voltage developed across R6 and R7 during this time passes through a one-half cycle of amplitude modulation.

d. Out-of-phase operation, first half-cycle. During half of each revolution of the acquisition antenna, the N-S resolver signal is 180° out of phase with the 4-kc carrier. When the signal at pin 4 of T1 is positive with respect to pin 6 of T1 and the signal at pin 6 of T2 is positive with respect to pin 4 of T2, the voltage induced in the secondary of T2 will cause current to flow from pin 4, to the negative terminal, through diodes V2B and V1A to pin 6 of T2. Pin 4 of T1 is positive with respect to pin 6 of T1. However, current cannot flow into pin 4 through V2A, as the cathode of that diode is connected to a more positive potential at pin 6 of T2. Therefore, current enters the secondary of T1 at the centertap and leaves at pin 6, passing through V1A to pin 6 of T2. The current path is completed with current flowing from pin 5 of T2 to ground through resistor R7 (developing a negative pulsation of voltage), and from ground through R6 to pin 5 of T1 (developing a positive pulsation of voltage).

e. Out-of-phase operation, second half-cycle. During the next half-cycle of the applied 4-kc signals, pin 6 of T1 will be positive with respect to pin 4 of T1, and pin 4 of T2 will be positive with respect to pin 6 of T2. The induced voltage across the secondary of T2 will cause current flow from pin 6 of T2 through diodes V2A and V1B to pin 4 of T2. The voltage induced in the secondary of T1 will cause current to leave pin 4 of T1 and pass through V1B to pin 4 of T2, and will pass to ground through R7 (again developing a negative pulsation of voltage). Current will then pass from ground through R6 to pin 5

of T1 (again developing a positive pulsation of voltage). Note that during the time when the resolver signal is out of phase with the 4-kc carrier, the signals developed across R6 and R7 are opposite in polarity to those developed when the resolver signal and the carrier are in phase.

f. Cathode followers V3B and V3A. The modulated pulsating voltage appearing across resistor R7 is applied to the grid of cathode follower V3B through the R-C filter consisting of R11, C3, R12, and C4. This filter removes the 8-kc component of the rectified signal so that the signal at the grid of the tube is a sine wave which varies about ground potential. This signal completes one cycle during each revolution of the acquisition antenna, and reaches the point of maximum positive value when the azimuth of the antenna is due north. Tube V3B conducts throughout the entire cycle, reproducing the applied sine wave across cathode resistor R13. This output signal is the north voltage. The amount of conduction of V3B when the applied signal is at maximum positive value is such as to develop a 30-volt potential at its cathode. When the applied signal is at maximum negative value, the conduction of the tube is such as to develop a potential of 3 volts at its cathode. As a result, the output signal is a positive voltage which varies sinusoidally within the limits of +3 volts and +30 volts. The voltage developed across R6 in the ring demodulator is at all times opposite in phase to that developed across R7. This signal is filtered by the R-C circuit, consisting of R8, C1, R9, and C2, and is applied to cathode follower V3A. The output of V3A is the south voltage. This signal is also a positive voltage which varies sinusoidally between +3 volts and +30 volts. However, the variation in this voltage is 180° out of phase with the variation in the north voltage.

g. Adjustment. The AMPLITUDE adjustment on each demodulator is used to adjust the length of the sweep on the PPI. These controls are set to fill the screen and obtain a round, calibrated display.

103. RANGE GATE CHANNEL (TM 9-5000-26, page 231)

a. General. The range gate channel is comprised of a gate multivibrator V1 and V2A, intensity limiter V2B, a clamper V3A, and a cathode follower V3B. These stages are located on the video amplifier chassis of the PPI. This channel produces two outputs, one used to unblank the PPI, the other used to aid in the generation of the sweep voltages. Tube V1 is a dual triode connected as a one-shot, cathode-coupled multivibrator with the V2A connected between the plate of one section and the grid of the other section of V1 for fast recovery. The multivibrator is triggered by the preknock pulse and produces 2 outputs, the sweep gate and the illuminating gate. The 2 outputs, taken from the same section of the multivibrator, are of the same polarity and duration, but differ in amplitude. The illuminating gate is the larger. The

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duration of the gate waveforms, which is determined by the position of the RANGE switch on the front panel of the PPI, may be equal to either 60,000 yards or 120,000 yards of radar range. The sweep gate is clamped at a fixed reference level by tube V3A and is applied through cathode follower V3B to the sweep generators, where it controls the width of the sweep voltages. The illuminating gate is clamped at a variable reference level by tube V2B and is applied to the control grid of the PPI tube to unblank the scope for the period of each sweep. The level at which tube V2B clamps the illuminating gate is determined by the setting of the INTENSITY control on the front panel of the PPI.

b. Gate multivibrator. The operation of the gate multivibrator is basically the same as the operation of a conventional one-shot, cathode-coupled multivibrator. The operation of the gate multivibrator is different in that a cathode follower is incorporated to reduce the recovery time. The gate multivibrator is capable of producing gate waveform durations of either 60,000 yards (395 microseconds) or 120,000 yards (735 microseconds). The gate multivibrator consists of tubes V1A, V1B, and V2A. Resistor R3 is the common cathode resistor. Coupling from the plate of V1A to the grid of V1B is through follower V2A, capacitor C3, or C4 and switch S1. Switch S1 is operated by the RANGE switch on the front panel of the PPI. In the quiescent state, tube V1B is conducting, its grid being returned to +250 volts through R7 and R8. The current through V1B flowing through resistor R3 raises the cathode to a potential sufficient to cut off tube V1A. The grid of V2A is connected directly to the plate of V1A. Since V1A is cut off, the potential at the grid of V2A is maximum. The preknock pulse is applied to the grid of V1A. The amplitude of the preknock pulse is sufficient to cause V1A to start conducting. The current through V1A causes a drop in potential at its plate. This drop is applied to the grid of V2A, causing a drop at the cathode of V2A. This drop is coupled through the capacitor, C3, to the grid of V1B, reducing the current through V1B. The decrease in current through V1B reduces the potential at the cathode of V1A. This increases the current flow through the A-section further reducing the potential of the plate. This action is cumulative and continues until the grid of V1B is driven below cutoff. With V1B cut off, only the current through V1A flows through R3. Since self-bias cannot cut off a tube, V1A continues to conduct even after the expiration of the preknock pulse at its grid. The grid of V1B is now at a negative potential. The instantaneous voltage coupled through C3 is slightly less than the initial change of voltage at the plate of V1A (due to the gain of the cathode follower being less than unity). Capacitor C3 now starts to discharge through R4, R7, and R8. As a result, the grid voltage of V1B starts to rise. The rate of rise is governed by the time constant of C3, R4, R7, and R8. The rate of rise can be adjusted by means of RANGE potentiometer R8. R8 is adjusted so that the sweep is actually 60,000 yards when in the 60,000-yard range position. When the grid potential of V1B reaches cutoff, V1B starts to conduct.

This causes an increase in the voltage drop across R3, increasing the bias on tube V1A. Greater bias on V1A produces a rise at its plate which is coupled through V2A and C3 to the grid of V1B. This switching action is also cumulative and results in the development of a cathode potential sufficient to cut off tube V1A. The return of V1A to a cutoff condition causes a sharp rise in potential at the grid of V1B. For correct operation of the circuit, it is necessary that the grid potential of V1B return to the quiescent-state value before the next preknock pulse arrives. In order for the grid potential to return to the proper value, capacitor C3 must regain the charge it lost during the 395-microsecond period that V1B was cut off. The recharging period of C3 is the recovery time of the gate multivibrator. The recharging path for C3 is from ground through R3 to V1B, from cathode to grid of V1B to one side of C3, from the other side of C3 through tube V2A to the power supply, and back to ground. The voltage at the plate of V1B is at a low value before the preknock pulse arrives; rises sharply as a result of the switching action initiated by the preknock pulse, remains at B-plus value for 395 microseconds (until the second switching action occurs), then drops sharply and stabilizes at the quiescent value. This results in the development of a positive rectangle waveform at the plate of tube V1B. The outputs from the gate multivibrator are taken from the plate circuit of V1B. One output, the illuminating gate, is developed across both R5 and R6 and is applied to tube V2B before application to the PPI. The other output, the sweep gate, is developed across only R6 and is applied to clamper V3A and cathode follower V3B before being transmitted to the sweep generators. The filament of tube V2 is connected to +150 volts to prevent an excessive difference in potential between the cathode and filament which might result in damage to the tube.

c. Sweep gate clamper and cathode follower. The sweep gate from the gate multivibrator is applied to tubes V3A and V3B to prepare it for application to the sweep generators. Stage V3A functions as a clamper, while V3B is a cathode follower. C1 couples the 75-volt square wave to the plate of diode V3A through R10. Since the cathode of V3A is returned to a voltage divider R11 and R23, the cathode is at +1 volt. When the positive square wave exceeds 1 volt, V3A clamps it at this 1-volt level. The result at the grid of V3B is a 75-volt square wave whose maximum positive value is 1 volt. Due to cutoff limiting, the effective signal at the grid of V3B is about 11 volts. V3B is a cathode follower with a gain of about 0.2. The result at the cathode of V3B is a positive square wave about 2 volts in amplitude. The cathode circuit of V3B appears in the sweep generator. The cathode resistors are R7, R8, and R9. The signal then that is applied to the sweep generators is about +2 volts. This signal, however, is large enough to cut off the diodes in the sweep generators.

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d. Intensity limiter (TM 9-5000-25 sheet 26). Tube V2B, in conjunction with the INTENSITY control, provides a means of controlling the amplitude of the illuminating gate which is applied to the control grid of the PPI tube. Resistors R3, R6, R7, and R8 are located just to the rear of the INTENSITY control knob, which is on the front panel of the PPI. The voltage at the cathode of tube V2B is determined by the INTENSITY control and may be varied within the approximate limits of +35 volts and +100 volts. The voltage at the plate of V2B (ST 44-161-2y, page 231) is obtained from the junction of R14 and R15 and during the time between illuminating gates, is approximately 80 percent of the voltage at the cathode of V2B. Therefore, during the periods between illuminating gates, tube V2B is cut off. When the positive illuminating gate is coupled through C6, the plate potential of V2B exceeds the cathode potential and the tube conducts. The current through R17 limits the potential to which the plate can rise to the value of the voltage existing at the cathode. The control grid of the PPI tube is connected directly to the plate of V2B. When the INTENSITY control is correctly adjusted, the limited illuminating gate appearing at the plate of V2B is of sufficient amplitude to raise the control grid potential of the PPI tube to a value just below visual cutoff. Application of negative signals to the cathode of the PPI tube at this time causes conduction, and only these signals are visible on the scope. Capacitor C5B maintains the d-c potential at the cathode of V2B constant. Contacts 3 and 4 of K1 are closed when the AZIMUTH LINE ring switch is depressed. This shorts out resistor R3, raising the potential at the cathode of V2B. This causes intensification of the sweep, which then appears at the steady steerable azimuth line. Switch S13 represents the ACQUISITION AZIMUTH SCAN switch, which in the CFF position grounds the junction of R7 and R8. This action reduces the intensity control voltage to such an extent that the scopes are not burned with the sweep is not rotating.

104. SWEEP GENERATORS (TM 9-5000-26, page 230)

a. General. There are 4 sweep generators for each PPI. The amplitude attained by the 4 sweep voltages is determined by the amplitude of the north, south, east, and west voltages produced by the ring demodulators and applied to the 4 sweep generators. These voltages vary sinusoidally as the antenna rotates. Therefore, the amplitude of the sweep voltage generated in each sweep generator will also vary sinusoidally as the antenna rotates. The relative amplitude of the 4 sweep voltages at any instant determines the direction in which the sweep on the PPI tube will be deflected. The sweep gate, generated in the range gate channel 1,000 times each second, is applied to the 4 sweep generators. This sweep gate has a duration equivalent to either 60,000 yards or 120,000 yards. The approximate duration of this signal, therefore, may be 395 or 735 microseconds. In either position of the RANGE switch, however, the physical length of the PPI sweep is the same, approximately 5 inches. The maximum amplitude achieved by the sweep voltages must be such as to produce

a deflection or sweep measuring approximately 5 inches. The signal amplitude necessary to obtain a sweep of this physical length is the same in both positions of the RANGE switch. In the 60,000-yard position of the switch, this signal amplitude must be attained in 395 microseconds, and the linear increase in the amplitude of the sweep voltage must be steep. In the 120,000-yard position of the switch, this same signal amplitude is achieved in 735 microseconds, and the slope of the voltage increase is more moderate. The movement of the electron beam of the PPI tube from the center of the screen to the edge must be at a uniform rate of speed in order for the sweep to be linear. This uniform movement is obtained by applying to the deflection coils signals which will produce a linear increase in current flow through the coils. The inductance of the coil winding must be overcome in order to obtain a linear increase in current. Hence, the voltage waveforms applied to the coils are trapezoidal. These signals are initially developed in the sweep generators.

b. Circuit construction. As the 4 sweep generators are identical, only the one which generates the north sweep voltage will be discussed. That stage contains tube V2A (pins 2 and 5), one-half of a dual diode. The plate of the diode is returned through approximately one megohm of resistance to the potential of the north voltage applied from the N-S ring demodulator. This voltage varies sinusoidally between +3 volts and +30 volts as the antenna rotates. The cathode of V2A, as well as of the other three diodes, is returned to ground through 3 series resistors which have a total value of 230 ohms. The plate of the diode is connected through contacts of the RANGE switch to one of 2 capacitors C3 or C7. The other side of the capacitor is connected through other contacts of the switch to one of 2 points in the resistive network in the cathode circuit. This point may be 5 ohms above ground or 10 ohms above ground. In the 120,000-yard position of the RANGE switch, the largest capacitor is in the circuit, and the capacitor is connected at the 10-ohm point in the cathode circuit.

c. Operation. In the absence of a sweep gate, the stage is conducting. Current flows from ground through the small value of the cathode resistance, through the diode, and through the very large resistance in the plate to a potential between +3 volts and 30 volts. Because of the distribution of the resistance in this current path, practically all of the voltage is dropped across the large resistors in the plate circuit. As a result, the capacitor connected between the plate and cathode has only a very small charge, and the d-c potential at the plate of the diode is only a small fraction of a volt above ground. Assume that the range switch is in the 60,000-yard position. Application of the positive 2-volt sweep gate to the cathode cuts off the diode as the charge on capacitor C3 holds the plate of the diode near ground potential. With the diode cut off, capacitor C3 begins to charge toward the potential of the applied north voltage. The charge path is from the capacitor through R5 and R2A, through

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V3B in the N-S ring demodulator, to ground through the +250-volt power supply, from ground through resistor R7, and to the other plate of the capacitor. The time constant of this circuit is approximately 6,000 microseconds. However, with the range switch in the 60,000-yard position, the duration of the sweep gate is only 395 microseconds. The capacitor is allowed to charge for about 6 percent of a time constant. During that period it will charge to a potential equal to about one-twentieth of the applied north voltage. The maximum amplitude of the sawtooth component of the signal varies between one-sixth of a volt and 1.5 volts. Because the capacitor is allowed to charge for only a small part of one time constant, the increase in potential which appears at the plate of the diode during the charge period is very linear. The desired trapezoidal waveform contains a rectangular component. This rectangular component appears in the output signal at the plate of the diode, primarily because of voltage divider action in the cathode circuit. A portion of the positive 2-volt sweep gate applied to the cathode of the stage will appear across the 3 small resistors in the cathode circuit and will raise the potential at that plate of the capacitor which is connected to the cathode circuit. This small rectangular signal is coupled through the capacitor to the plate of the diode. The subsequent charging of the capacitor causes the sawtooth component of the signal to be superimposed on this rectangular component. A larger capacitor is introduced into the circuit when the RANGE switch is in the 120,000-yard position. The larger capacitance results in a longer time constant in the charge of the capacitor. Therefore, although the time during which the capacitor may charge is increased, the voltage attained by the capacitor during the charge period remains approximately 5 percent of the applied north voltage. As the north voltage varies with the rotation of the antenna, the amplitude of the generated trapezoidal north sweep voltage also varies with the rotation of the antenna.

d. Adjustment. The sweep generator chassis contains a HORIZONTAL BALANCE and a VERTICAL BALANCE adjustment; these controls are used to cause the east and west as well as the north and south outputs to cancel when no signal is present. To adjust these controls, turn off the two BALANCE switches on the demodulator chassis and adjust the balance controls until a sharp spot is obtained.

105. SWEEP CHANNEL AMPLIFIERS (TM 9-5000-26, page 232)

a. General. The purpose of the sweep channel amplifiers is to amplify the trapezoidal outputs of the sweep generators and cause a sawtooth of current to flow through the deflection coils. The south section of voltage amplifiers V1 and V2, clamper V3A, and power amplifier V4. In order to insure that the sweep will always start at the same point on the PPI, the signal at the grid of power amplifier V4 is maintained at a constant reference level

by clamper V3A. Since it is current and not voltage which is of importance in an electromagnetic deflection system, current feedback is employed from V4 to V1 to insure that a linear output current waveform is produced. In the north section, V5 and V6 are the voltage amplifiers, V3B is the clamper, and V7 is the power amplifier. The current feedback is from V7 to V5. The plate load for each power amplifier is one of the 4 windings of the deflection coil of the PPI tube. Since the N-S sweep channel is identical to the E-W sweep channel, only the N-S sweep channel will be discussed.

b. Voltage amplifiers. The positive trapezoidal sawtooth voltage from the sweep generator is applied to the control grid of V1. The cathode of V1 is returned to ground through R14 in series with R16, the cathode resistor of tube V4. This provides degenerative feedback from tube V4 to V1. The positive signal at the grid of V1 results in a positive signal at the grid of V4. Since R16 is unbypassed, the cathode of V4 rises with the grid. Because R16 is common to both V4 and V1, the cathode potential of V1 also rises. Therefore, the grid-to-cathode signal of tube V1 is the grid-to-ground signal minus the feedback signal from R16. The feedback improves the waveform at the output of V4 by reducing the distortion which may be introduced by the amplifiers. The signal which is applied to the grid of V1 is amplified by a certain factor in the output, but the distortion caused by the nonlinear characteristics of the tube is not amplified in the output. This is also true with respect to both V2 and V4. Consequently, when the signal across R16 is fed back so as to oppose the grid signal of V1, the distortion component of the degenerative voltage is amplified by the same factor as the signal on the grid of V1. As a result, the current which is drawn through the deflection coil by V4 is, as near as possible, a linear sawtooth current. Resistor R14 provides self-bias for tube V1. The value of C6 is such as to bypass the initial rise of the trapezoidal waveform around R14. This causes a sharp spike at the leading edge of the sawtooth waveform. The purpose of the spike is to overcome the initial reactance of the deflection coil to the rising sweep current. C2 and R10 reduce the high-frequency response of V1, thus preventing the stage from being shocked into high-frequency oscillations by the input signal. The negative output signal from V1 is coupled by C1 and R11 to amplifier V2. Tube V2 is a conventional voltage amplifier. Its positive output is coupled through C5, clamped at a negative reference level by the action of tube V3A, and applied to the control grid of V4.

c. Clamper. Tube V3A is a cathode follower which operates to maintain a constant d-c reference level at the control grid of tube V4. A constant d-c level at the grid of V4 is necessary in order to cause the electron beam to start from the same point on the PPI for each successive sweep. The grid of V3A is held at approximately -65 volts by voltage divider R6-R12. The cathode potential, which is determined by the current flow through V3A and through

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R5, is slightly less negative than the grid. The voltage is -55 volts at the cathode. The cathode of tube V3A is connected through resistor R9 to the control grid of V4. Resistor R9 will suppress parasitic oscillations. Assume that the d-c potential at the grid of V4 becomes slightly more positive than normal. The cathode of V3A would become more positive, reducing the current through V3A. This would reduce the current through R5, decreasing the voltage drop across it. Less voltage dropped across R5 would cause the potential at the cathode of V3A and at the control grid of V4 to become more negative, thus returning the potential at the grid of V4 to its original value. C3C bypasses R12, thereby maintaining the d-c potential at the grid of V3A at a constant value. The negative 65-volt potential obtained from voltage divider R6-R12 is used in similar fashion in the north section as well as in the E-W sweep channel.

d. Power amplifier. During the period between sweep signals, tube V4 is cut off. The plate load for tube V4 is one-half of either the north-south or the east-west deflection coil. The function of V4 is to deliver a linear sawtooth current through this plate load. Not all of the current through R16 flows through the deflection coil, as some of this current is drawn by the screen grid of V4. The feedback voltage used to cause the current waveform at the plate of V4 is one-half of either the north-south or the east-west deflection coil. The function of V4 is to deliver a linear sawtooth current through this plate load. Not all of the current through R16 flows through the deflection coil, as some of this current is drawn by the screen grid of V4. The feedback voltage is used to cause the current waveform at the plate of V4 to be as linear as the signal at the grid of V1. Therefore, the effect of the screen grid current of V4 on the feedback signal must be canceled. This is accomplished by using the screen grid current to develop a signal which is 180° out of phase with the feedback signal and algebraically adding it to the feedback signal. When the trapezoidal waveform appears at the control grid of V4, the tube conducts and screen grid current is drawn through R7. The current through R7 develops a negative-going signal at the junction of R7 and R13. This signal is applied to the feedback circuit through R13 and isolating resistor and cancels that part of the feedback signal which results from the flow of screen grid current through it.

106. PPI VIDEO AMPLIFIER (TM 9-5000-26, page 231)

a. General. The input to the video amplifier is the positive video and mark output of the video and mark channel. Tube V4 is a conventional video amplifier. An inductor is included as part of the plate load in order to maintain the shape of the video and marks. The reference level of the video and mark signal input is kept constant by crystal clamper CR1. The negative output of tube V4 is applied to the cathode of the PPI tube. These negative

signals intensify the sweep and appear on the scope as brightened portions of the PPI sweep.

b. Operation. The video and mark signals from the video and mark channel are applied through contacts 1 and 9 of K1 and through the GAIN potentiometer (located on the rear of the front panel of the PPI) to the control grid of video amplifier V4. The signals, which must be amplified by V4, carry greatly in duration. The acquisition azimuth mark is a 735-microsecond signal, while the range mark signals are only 0.5 microsecond in duration. Therefore, both high-frequency and low-frequency compensation must be employed in the circuitry of V4. The principal cause of the reduced gain of the high-frequency components of a signal is the effective reduction of the plate load caused by the shunting effect of the interelectrode and distributed capacitances. This effect is compensated for by inductance L1, which is in series with R21. Coil L1 serves to boost the response at high frequencies, but has practically no effect upon the low-frequency response. The low-frequency response of a video amplifier is affected by the grid coupling circuit. Low-frequency components are attenuated by the coupling capacitor because of its increased reactance at the lower frequencies. Coupling capacitor C8 and developing resistor R19 as a voltage divider will have a frequency decrease as the reactance of C8 increases. Hence, a smaller percentage of the low-frequency components of a signal is developed across R19 and applied to V4. Capacitor C8 is made as large as is reasonably possible in order to compensate for the attenuation of the low-frequency components. In many applications video amplifiers operate with self-bias obtained from an R-C network in the cathode circuit. This type of biasing method causes low-frequency attenuation. In tube V4, fixed bias is obtained from voltage divider R18-R20, thus eliminating the need for the R-C network in the cathode circuit. The third low-frequency compensating device in the video amplifier stage is the filter network, C54-R12, in the plate circuit. Capacitor C7 in the grid circuit maintains a constant d-c potential at the junction of R19 and R20 and prevents R18 from being a part of the grid-developing resistance of V4. The signal at the grid of V4 is positively clamped at a -2-volt CR1. The purpose of CR1 is to prevent variations in the intensity of the signals displayed on the PPI. When a strong signal of long duration is applied, capacitor C8 is allowed to charge. For a period following the termination of the signal, the potential of the grid of V4, in the absence of CR1, would be of a comparatively high negative value. This negative potential would prevent a signal immediately following the large signal from appearing on the PPI at its normal intensity. However, any negative potential in excess of -2 volts is shorted to ground through CR1, and a reference level of -2 volts is maintained at the control grid of V4.

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107. PPI TUBE (TM 9-5000-26 pages 227, 228)

a. General. The PPI tube is a 10-inch, high-persistency, cathode ray tube. The tube is of the electromagnetic deflection type, utilizing stationary deflection coils. The fluorescent material on the screen of the tube is a compound which has the characteristic of producing blue light. The electron gun of the PPI tube is conventional, being made up of an indirectly heated cathode, a control grid, and the first accelerating anode. The cathode-to-control grid potential is such that the tube operates at visual cutoff, conducting only when the positive illuminating gate is present on the grid and negative video or mark signals are also present at the cathode. The second accelerating anode is a graphite coating on the inside of the glass envelope known as aquadag. The coating extends from near the focus magnet to the fluorescent screen of the tube. During operation, the aquadag is at a positive potential of 8,500 volts. In addition to accelerating the electrons, the aquadag collects the stray electrons which result from secondary emission at the fluorescent screen. The deflection coils are wound on an iron yoke which is placed around the neck of the PPI tube. The electron beam is deflected by passing through the magnetic field resulting from the sawtooth current flow through the coils. Focusing of the beam is accomplished by a permanent magnet placed around the neck of the PPI tube in such a manner that it can be moved axially along the tube. The beam is centered by tilting the focus magnet.

b. Operation. The cathode of the PPI tube is a small cylinder with an oxide coating at one end. The oxide coating permits great emission from the cathode in the desired direction. The cathode is indirectly heated by a filament coil inside the cathode cylinder. To prevent the filament voltage from affecting the electron beam, the filament coil is wound in a double spiral so that the magnetic field of half the winding is canceled by the equal but opposite magnetic field of the other half. The control grid is a metal cylinder placed around the cathode. The end of the grid cylinder is covered by a disk with a small hole in the center, through which the electrons are allowed to pass. When the grid-to-cathode potential is varied, either a greater or lesser number of electrons escape through the hole, thereby increasing or decreasing the brightness of the trace seen on the screen. Two devices accelerate the electrons from the cathode to the screen. The first anode, which operates at a positive potential of 320 volts, is located just ahead of the control grid. The second accelerating anode is the aquadag coating, which operates at a potential of +8,500 volts. In addition to accelerating the electrons, the aquadag attracts the secondary electrons emitted from the screen, thereby completing the circuit for the flow of electrons from the cathode. The number of secondary electrons reaching the aquadag must equal the number that originally strike the screen, otherwise the screen would acquire a large negative voltage which would interfere with the normal operation of the tube. The velocity of the electrons traveling

from the cathode to the screen is so great that they are not attracted directly to the aquadag. Although the control grid narrows the electron beam, it is not capable of appreciable focusing action. Focusing is accomplished by the field of a permanent magnet. As the electrons pass through the focusing field, they are acted on in such a way as to cause them to describe a spiral path like the tread on a wood screw as they proceed to the screen. The electrons that are far off the axis of the tube when they enter the focusing field are made to follow a path that will make them meet the electrons that come into the field almost parallel with the axis of the tube. The focus magnet is tilted to cause the meeting point to be at the center of the screen. Deflection of the beam is accomplished by the varying magnetic fields of the deflection coils. The four magnetic fields are the product of the simultaneous application of the 4 inphase sweep voltages to the deflection coils. The outward deflection of the beam, which produces the radial sweep, is caused by the sawtooth currents. Rotation of the sweep is caused by the sinusoidal variation in the amplitude of the sweep voltages.

108. PPI HIGH-VOLTAGE POWER SUPPLY (TM 9-5000-26, page 208)

a. General. The high-voltage power supply supplies +5,000 volts for the precision indicator. The power supply and indicators are located (physically) in the battery control console.

b. Circuit analysis. The +8,500-volt and +5,000-volt outputs are obtained from a voltage doubler circuit consisting of transformer T1, capacitor C1, and diode rectifiers V1 and V2. The 120-volt, 400-cycle power is applied to the primary of transformer T1, which steps this voltage up to 3,300 volts between terminals 3 and 5 (secondary). The cascade-arranged voltage doubler causes capacitor C1 to remain charged to the peak voltage. Therefore, twice the peak voltage appears on the plate of the rectifier V2 which, by maintaining a charge on capacitor C2, produces the +8,500 volts of direct current. Capacitors C2 and C3 and resistors R1, R2, and R3 comprise a capacitive-type filter and voltage divider network. The +8,500 volts output for the PPI electron collector is obtained at jack J2, and the +5,000 volts for the precision indicator electron collector at jack J3. Very little current is drawn from this power supply. To obtain a more detailed description of the circuit operation of this power supply, see BTL vol V, sec 9-III-6.

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CHAPTER 7

TARGET DESIGNATION

Section I. TARGET DESIGNATE BLOCK DIAGRAM

109. GENERAL

The purpose of the acquisition radar is to search the area around the battery for enemy targets. When the acquisition radar displays a target, some means must be available to mark or designate this target in range and azimuth so that the track radar may be positioned to this target quickly and accurately. The device that accomplishes this mission is the target designator. The target designator is made up of two sections, the designator control system and the designator presentation system. The job of the designator control system is to generate a flashing azimuth line and a range circle for designating the target in azimuth and range. The designate control system also generates the electronic cross which shows the azimuth and range of the track radar. The designate presentation system (fig 2) consists of 2 precision indicators. These indicators display an expanded view of the area centered around the intersection of the range circle and the flashing azimuth line or the area around the electronic cross and enable the operators to designate targets more accurately and also to see the target more clearly.

110. TARGET DESIGNATE COMPLETE BLOCK (TM 9-5000-26, page 223 and TM 9-5000-25, sheets 24 and 25)

a. Designation marks. When an enemy target is seen on the PPI at the tactical control console, the operator immediately depresses the azimuth line control and rotates the steady steerable azimuth line to the azimuth of the target. At the same time he turns the range circle control until the range dot is over the target. In performing these operations, the azimuth resolver channel develops a signal which is sent to the PPI to form a steady sweep. This control at the same time also controls signals which go to the acquisition mark generator channel to form the acquisition azimuth mark (QAMK). The QAMK is sent to the mixer channel where it is added to the other marks to be displayed on the PPI as a flashing azimuth line. When the range circle control is turned, the acquisition range control channel adjusts the voltage at the input to the acquisition range channel. This channel produces the acquisition range mark (QRMK) which is sent to the mixer to be displayed as the range circle.

b. Electronic cross. This element of the display is produced by 4 signals generated in circuits associated with the target-tracking radar. Its purpose is to indicate the position of the target being tracked by the target-tracking radar. This is accomplished by causing the electronic cross to be generated at an azimuth and range which correspond to the settings of the azimuth and range servo systems of the target-tracking radar. The electronic cross consists of a 5,000-yard radial line and a 10° arc. The azimuth of the electronic cross is determined by the track azimuth resolver which is connected to the track antenna. This resolver's output is sent to the track mark generator channel where the tracking azimuth mark (TAMK) and the tracking azimuth gate (TAGA) are generated. The tracking azimuth mark is sent to the mixer channel where it is gated by the tracking range gate (TRGA) to form the 5,000-yard radial line. The tracking azimuth gate is also sent to the mixer channel where it gates the acquisition-tracking range mark (QTRMK) into a 10° arc. The QTRMK then forms the arc of the cross.

c. Designate indicator system. The designate indicator system consists of 2 precision indicators and associated circuits, one at the target console and one at the battery control console. The precision indicators provide a means of locating a target accurately with respect to the azimuth and range marks, once the target appears on the PPI display. One precision indicator displays a 30° by 5,000-yard sector of the PPI display and the other a 30° by 8,000-yard sector. The sector displayed by a precision indicator is determined by the position it occupies in the system, either at the battery control console or at the target console. If in the battery control console, the sector is centered at the intersection of the flashing steerable azimuth line and the range circle. If in the target console, the sector is centered about the electronic cross. The precision indicators have a B-type presentation, employing 5-inch cathode ray tubes and associated magnetic deflection systems. Each designate indicator system contains a range sweep channel, an azimuth sweep channel, a PI video amplifier, an unblanking channel, and front panel controls. The range sweep channel in the battery control console receives the QTRGA and the one in the target console receives the TRGA. The TRACK-ACQ switch is not wired for the SQ signals to interchange. From this input the range sweep channel generates the 5,000- or 8,000-yard range sweep which is centered at the range setting of either the tracking range system or the acquisition range system. The azimuth sweep generator in each system receives an input from the azimuth resolver of the corresponding system depending on the location of the indicator (radar control trailer, or the battery control trailer) as to signals available to it. From this input, it produces a 30° azimuth sweep which is centered upon either the radial line or the flashing azimuth line. The precision video amplifiers receive the output from the video and mark channel and apply these signals to the cathodes of the PI's. The unblanking signals received









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by the unblanking channel are the tracking range gate with the tracking azimuth gate or the acquisition range gate with the acquisition azimuth gate. Selection of the unblanking voltages is accomplished by the signal distribution wiring plan shown in TM 9-5000-25, sheet 24. The recurrence frequency and time of occurrence of the input signals to the precision indicators are such that the display occurs once for each revolution of the acquisition antenna. The sweep moves across the tubes in azimuth in synchronism with the movement of the acquisition antenna. The rapidity of movement of the sweep is determined by the rate of rotation of the acquisition antenna. If no range sweep voltages were applied, the presence of a permanent magnet would cause the electron beam to move straight across the bottom of the tube. However, with the application of the range sweep voltages, the beam is deflected upward from 170 to 500 times during the period of one 30° azimuth sweep. The number of range sweeps for each azimuth sweep is determined by the rate of rotation of the acquisition antenna. The precision indicator is intensity modulated with the video and marks from the video and mark channel through the PI video amplifiers, and either the electronic cross or the flashing steerable azimuth line and range circle will be always displayed. If the range and azimuth of both radars coincide, all of these marks will be visible on the precision indicator.

d. Chart of signals. The range circle, the flashing steerable azimuth line, the electronic cross, and the video signals are all made to appear on the indicators through intensity modulation of the individual tubes. A single recurring waveform produces the range circle, and another waveform produces the flashing steerable azimuth line. These two signals are generated in circuits of the acquisition radar. However, 4 separate signals are required to produce the electronic cross, and all 4 of these signals are generated in circuits controlled by the target-tracking radar. Table III lists the various signals which are used to produce the desired display on the PPI's and PI's.

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Table III. PPI and PI signals.

NAME	SYMBOL	WAVEFORM	DESCRIPTION	USE	GENERATED
Acquisition range mark	QRMK		A positive pulse of 0.5-microsecond duration, generated once each sweep at a range determined by the RANGE CIRCLE handwheels.	Intensifies a point on each sweep to form the range circle. Used at both PPI's and the PI in the battery control console.	Acquisition range channel.
Acquisition range gate	QRGA		A positive square wave 48 microseconds in duration, generated once each sweep. Leading edge occurs 15 microseconds before appearance of QRMK.	Aids in unblanking the precision indicator and gates the range sweep channel of the battery control console's precision indicator to generate the expanded range sweep.	Acquisition range channel.
Acquisition azimuth mark	QAMK		A positive square wave, 612 microseconds in duration, generated once per revolution of the acquisition antenna at an azimuth determined by the AZIMUTH LINE controls.	Intensifies a single sweep on the acquisition indication on the battery control console to produce the flashing steerable azimuth line.	Acquisition mark generator channel.
Acquisition azimuth gate	QAGA		A positive waveform with a duration equal to 30° of rotation of the acquisition antenna. Appears once per revolution of the acquisition antenna at an azimuth determined by the AZIMUTH LINE controls.	Aids in unblanking the precision indicator at the battery control console.	Acquisition mark generator channel.
Acquisition tracking range mark	QTRMK		A positive pulse of 0.5-microsecond duration, generated once each sweep at a range determined by the range setting of the tracking radar.	When gated by a 10° portion of the TAGA, the resultant signal will produce the arc of the electronic cross.	QTRMK channel of the range tracking system.
Tracking azimuth gate	TAGA		A positive waveform with a duration equal to 30° of rotation of the acquisition antenna. Appears once per revolution of the acquisition antenna at the azimuth position of the tracking radar.	A 10° portion of this signal gates the QTRMK to produce the arc of the electronic cross. The full 30° width aids in unblanking the precision indicators for the TRACK display.	Track mark generator.
Tracking azimuth mark	TAMK		A positive 610-microsecond square wave generated once per revolution of the acquisition antenna at the azimuth position of the tracking radar.	The coincidence of this signal with the TRGA produces the radial line of the electronic cross.	Track mark generator.
Tracking range gate	TRGA		A positive 30-microsecond square wave generated once each sweep. The leading edge appears 15 microseconds before the QTRMK.	Coincidence with the TAMK produces the radial line of the electronic cross. Aids in unblanking the precision indicators and gates the range sweep channel of the battery control presentation system for the TRACK display.	Track range gate cathode follower of the range tracking system.

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Section II. RESOLVERS

111. LINE SLEW RESOLVERS (TM 9-5000-25, sheets 25 and 26)

a. General. When the position of the steerable azimuth line is changed, the ring depress switch at the target designate control panel is normally operated. As a result, the sweep on the PPI tube will no longer rotate in synchronism with the rotation of the acquisition antenna. Instead, it is deflected in a direction determined by 2 line slew resolvers. The rotors of these resolvers are controlled by the AZIMUTH LINE control knob, which thus controls the sweep in its deflection. The resulting sweep appearing on the indicators is the steady steerable azimuth line. Target designation in azimuth is accomplished by use of the line slew resolver and its associated switching circuit (fig 29). Line slew resolver B2 is located in the target designate control panel and has applied to one of its stator windings a 4-kc signal from the 4-kc oscillator. The rotor of the line slew resolver is mechanically coupled to the ACQ line resolver B4, the ACQ control transformer B5, and drive-motor tachometer B6. The output voltage from the line slew resolver is taken from the rotor windings of the control transformer B5 located in the target designate control panel. This output voltage is available to both PPI's. When the outer ring of the AZIMUTH LINE control is depressed, the video and marks are removed from the PPI tube at the battery control console, and the acquisition range mark is reapplied. The sweep is brightened, and the input to the acquisition presentation system (PPI) normally obtained from the azimuth channel amplifier is replaced by the output of the line slew resolver. This means that the sweep instead of rotating in synchronism with the acquisition antenna will remain stationary except when the AZIMUTH LINE control is moved. This sweep, which is brighter than normal, is the steady steerable azimuth line. Since rotors of the line slew resolver and the AZIMUTH LINE resolver in each console are mechanically connected, the displacement angle of the 2 line slew resolvers and the displacement angle of the 2 azimuth line resolvers are equal. Therefore, when the outer ring of the AZIMUTH LINE control is released, the flashing steerable azimuth line generated from the output of the azimuth line resolver will appear at the same azimuth at which the steady steerable azimuth line appeared while the outer ring was depressed.

b. Relay circuit (TM 9-5000-26, page 223, TM 9-5000-25, sheet 26). Associated with the AZIMUTH LINE control knob is a ring depress switch. Operation of the switch at the battery control console has no effect upon the presentation at the target console. The switching circuit actuated by K1 ring depress relay substitutes acquisition range mark for acquisition video only in the PPI at the battery control console. When it is desired to position the steerable azimuth line, the ring depress switch at the target designate control panel may be operated, and the associated control knob adjusted. While the

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switch is held closed, -28 volts energize relay K1 in the PPI at the battery control console (TM 9-5000-25, pages 227 and 228). As a result, mixed video and marks are no longer applied to the PPI at the battery control console, and only the acquisition range mark is applied. A second effect is the brightening of the PPI display. Normally, the intensity is adjusted so that the tube operates very close to visual cutoff. Relay K1 increases the intensity of the sweep in order that the steerable azimuth line will be plainly visible. In the intensity control circuits a filter has been added to eliminate the splatter effect sometimes seen when either AZIMUTH LINE switch is operated. The circuits consisting of R13-C2A and R14-C2B are connected through contacts of relay K1 to the centertap of the INTENSITY control. Upon operation of the AZIMUTH LINE switch, the increase in intensity is delayed by the time required for C2B to charge. Upon release of the switch, the decrease in intensity is delayed by the time required for C2A to charge. This more gradual voltage change eliminates the splatter effect. Resistors R13 and R14 provide a discharge path for their associated capacitors. When K1 is energized, the circuit through the coil of relay K1 in the target designate control panel is completed, and K1 is energized. In normal operation, the sweep deflection of the PPI tube is determined by the N-S and E-W voltage outputs of the azimuth channel. When K1 is energized, these voltages are no longer applied to the acquisition presentation system (PPI) sweep channel. In their place, the N-S and E-W voltages obtained from the line slew resolvers are applied. The direction in which the sweep is deflected may then be controlled by the line slew resolvers, and the sweep itself appears as the steady steerable azimuth line. The steerable azimuth line may also be positioned from the tactical control console. Operation of the ring depress switch S1 at the target designate control panel energizes relay K1 in the PPI at the battery control console. When energized, this relay intensifies the PPI sweep, removes the mixed video and marks from the PPI presentation, substitutes the acquisition range mark, and energizes relay K1. When energized, relay K1 permits the E-W and N-S voltage outputs of the line slew resolvers to be applied to the sweep channel of the acquisition presentation system (PPI) at the battery control console. Capacitors C1, C2, C3, and C4 provide the load for the line slew resolvers when neither relay is energized. They are selected to match the load which is presented when either relay is energized.

112. AZIMUTH LINE RESOLVERS (TM 9-5000-25, sheet 25)

a. General. Also connected to the AZIMUTH LINE control knobs are the rotors of the 2 azimuth line resolvers. As the rotor of the line slew resolver is moved, the rotors of the azimuth line resolvers move correspondingly. The outputs of the azimuth channel are applied to the azimuth line resolvers. The connections are such that the sum of the displacement angles of the 2 azimuth line resolvers is subtracted from the angular displacement of the acquisition

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azimuth resolvers. At this instant, the E-W output voltage is passing through zero. This voltage null is detected in the acquisition mark generator channel and causes the generation of the acquisition azimuth mark, which appears on the indicators as the flashing steerable azimuth line.

b. Operation. There are two azimuth resolvers. The azimuth line resolver, R4 is located in the acquisition azimuth control unit in the target designate control panel. The track azimuth resolver, B4 is located in the azimuth data unit at the target-tracking antenna. The rotor of the azimuth line resolver is mechanically connected to the rotor of an associated line slew resolver B2. Control of the angular displacement of the rotor is accomplished by an AZIMUTH LINE control knob at the target designate control panel. The N-S and E-W output voltages of the azimuth channel are applied to the azimuth line resolver, B4, in the target designate control panel. B4 is mechanically coupled to line slew resolver, B2. The output voltages from B2 are applied to the other azimuth line resolver, B4, at the target-tracking antenna. The connections are such that the sum of the displacement angles of resolvers B2 and B4 is subtracted from the acquisition azimuth angle. The E-W and N-S outputs from resolver B4 are applied to the acquisition mark generator channel which detects the null voltage that occurs when the acquisition azimuth equals the sum of the displacement angles of the 2 azimuth line resolvers. This null causes the generation of the acquisition azimuth mark and gate, and hence the appearance on the scopes of the flashing steerable azimuth line. In addition, the E-W voltage output of the azimuth line resolvers is applied to the azimuth sweep channel of the designator presentation system when the TRACK-ACQ switch is in the ACQ position.

113. TRACK AZIMUTH RESOLVER

The rotor of the track azimuth resolver, B4, is mechanically connected to, and is rotated by, the tracking antenna. The rotor output voltages are proportional to the sine and cosine of the angular azimuth difference of the acquisition and target-tracking radars. These output voltages are applied to the tracking mark generator channel for use in the production of the tracking azimuth gate and the track azimuth mark. These signals are generated when the azimuth of the acquisition antenna is the same as the azimuth of track antenna. The E-W output voltage is also applied to the azimuth sweep channel of the designator presentation system. When the TRACK-ACQ switch is placed in the TRACK position, the signal is used to produce a 30° azimuth sweep for that precision indicator. This 30° sweep is centered at the azimuth of the tracking antenna. This resolver is located in the azimuth data unit at the target antenna trailer and is identical to the acquisition azimuth resolver. The N-S and E-W voltage outputs of the azimuth channel are applied to its stator windings. The circuit connections are such that the output voltages indicate an angle equal to the

acquisition azimuth minus the tracking azimuth plus 180° . When the two radars are at the same azimuth, the resultant output voltages indicate an angle of 180° . As the sine function of an angle of 180° is zero, the E-W output voltage from the tracking azimuth resolver at that instant is zero. The output voltages are applied to the tracking mark generator channel, which detects this null voltage. In addition, the E-W voltage output is applied to the azimuth sweep channel of the designator presentation system to develop the azimuth sweep of the precision indicator when the TRACK-ACQ switch is in the TRACK position.

Section III. MARK GENERATOR CHANNELS

114. INTRODUCTION (TM 9-5000-26, page 234)

a. General. All azimuth data used in the acquisition presentation system are initially supplied in the form of modulated voltages from the azimuth channel and the resolver channel. Excitation for the resolvers is furnished by the 4-kc oscillator channel, which also produces an unmodulated carrier signal. This 4-kc carrier is used in the demodulation of the output voltages of the azimuth channel and the resolver channel. The proper generation of sweeps on the PPI and PI is dependent upon this demodulation. In addition, the 4-kc carrier is used in the mark generator channels together with modulated output voltages of the azimuth channel and the resolver channel. The mark generator channels produce the azimuth signals required to unblank the precision indicators, to generate the electronic cross, and to produce the flashing steerable azimuth line. Two identical and interchangeable mark generator chassis are used in the target designate control system. The operation of these units differs only in the source of their input signals and the use of their output signals. One of these chassis is physically located in each precision indicator unit. The precision indicator at the target console contains the tracking mark generator chassis. The acquisition mark generator chassis is located in the precision indicator unit at the battery control console.

b. Acquisition mark generator channel. The inputs to the acquisition mark generator channel are the sine (E-W) and cosine (N-S) outputs of the azimuth line resolvers in the resolver channel (fig 31, parts (1) and (6)), the 4-kc carrier, and the preknock pulse. The outputs from this unit are the acquisition azimuth mark (fig 31, part (4)) and the acquisition azimuth gate (fig 31, part (11)). The characteristics of these signals are the same as the tracking azimuth mark and gate, except for their time of occurrence. These signals occur once each revolution of the acquisition antenna at an azimuth determined by the azimuth line resolvers. The acquisition azimuth mark is applied to the mixer channel and produces the flashing steerable azimuth line. The acquisition azimuth gate is applied to the unblanking channel of the designator presentation system when the TRACK-ACQ switch is in the ACQ position.

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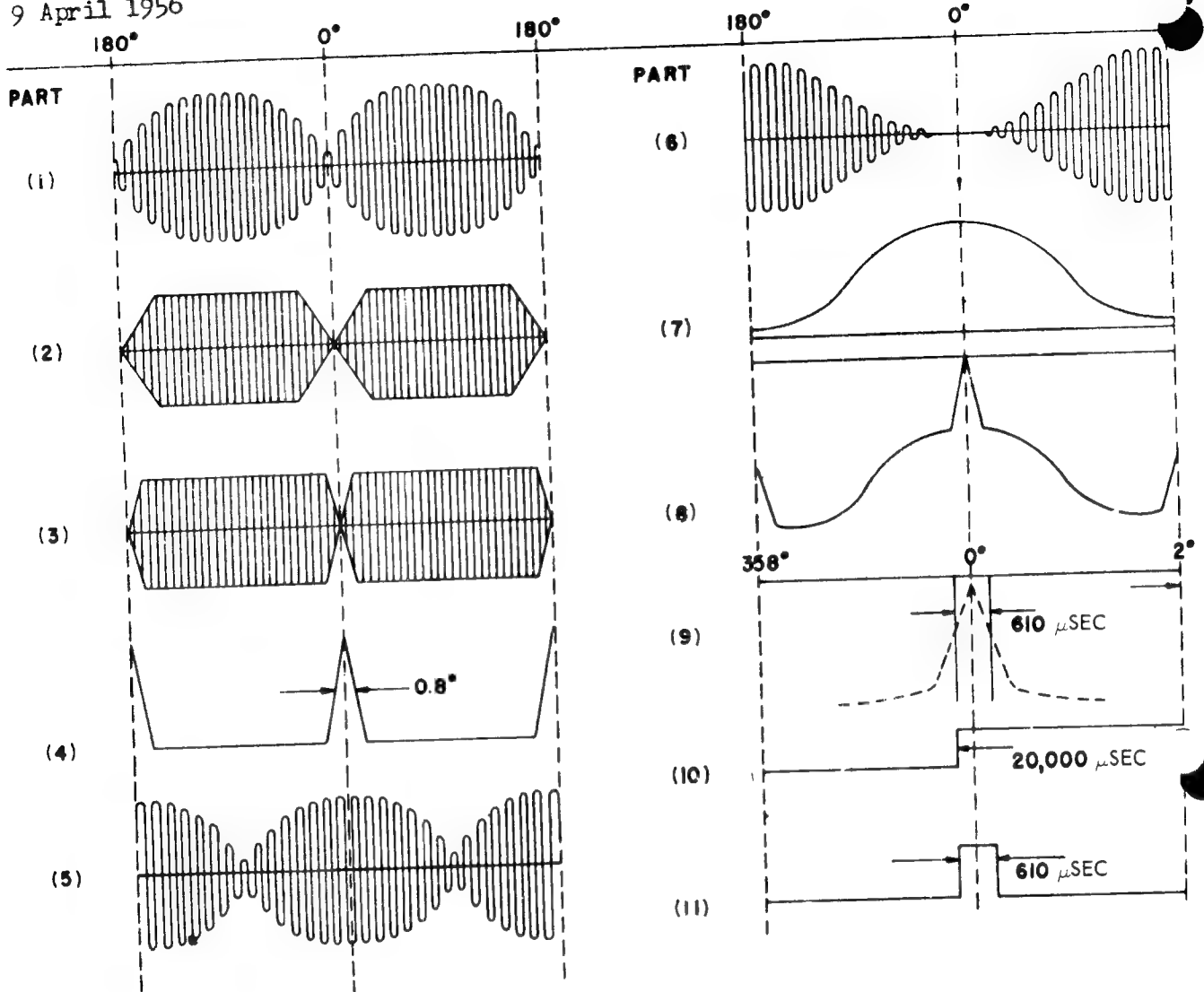


Figure 31. Mark generator waveshapes.

c. Tracking mark generator channel. The inputs to the tracking mark generator channel are the E-W and N-S outputs of the tracking azimuth resolver, the 4-kc carrier, and the preknock pulse. From these inputs are generated the tracking azimuth mark and the tracking azimuth gate. These signals occur once each revolution of the acquisition antenna, appearing at the azimuth of the tracking antenna. The tracking azimuth mark, a 610-microsecond waveform, is applied to the mixer channel for use in the production of the radial line of the electronic cross. The tracking azimuth gate, the duration of which is equal to 30° of rotation of the acquisition antenna, is applied to the mixer channel for use in the production of the arc of the electronic cross. The tracking azimuth gate is also used to unblank the precision indicator for the 30° azimuth sweep when the associated TRACK-ACQ switch is in the TRACK position.

115. 40° LIMITER V1A

The E-W output signal from either the azimuth line resolver or the tracking azimuth resolver is applied to V1A. The positive portions of the signal are limited by grid limiting. The negative portions of the signal cut off the tube. The output of V1A is a 4-kc signal with a modulated V-section corresponding to 40° of antenna rotation. The V-portion of the output from V1A marks where null points occur in the E-W signal input. There are 2 nulls per revolution of the acquisition antenna. Therefore, in the output of V1A, there are 2 modulated 40° V-sections per revolution of the antenna. The input to V1A in the acquisition mark generator channel is the modulated E-W voltage output of the azimuth line resolvers. In the tracking mark generator channel, the input to V1A is the E-W voltage output of the tracking azimuth resolver. Although the source of the input signal varies, the characteristics of the signal do not. This signal is a 4-kc voltage whose amplitude and phase vary sinusoidally as the acquisition antenna rotates. Two times during each revolution, the sine function equals zero. At these two instants, the input signal to V1A passes through a null. At the instant of each null, the 4-kc voltage, which is then of zero amplitude, experiences a phase reversal of 180°. The 40° limiter acts to limit the positive and negative alternations of the 4-kc signal to a fixed value. This is accomplished by grid limiting (resistor R1) and cut-off limiting. However, as the modulation of the 4-kc signal approaches, passes through, and recedes from each null, the amplitude of the signals is not sufficient to cause grid current flow or to cut off the tube, and limiting does not occur. Instead, the output waveform will reproduce the reduction and subsequent increase in amplitude of the 4-kc signal as the null is approached and passed. Tube characteristics and the maximum amplitude of the modulated 4-kc input signal are such that the modulation envelope is reproduced through a 40° arc of antenna rotation. If the amplitude of the excitation voltage generated in the 4-kc oscillator and applied to the acquisition azimuth resolver were increased, the maximum amplitude of the signal applied to V1A would also increase, and the duration of the 40° V-section would be reduced. Two 40° V-sections appear in the output of V1A during each revolution of the acquisition antenna.

116. MIXER AMPLIFIER V3A

The null which occurs in the E-W resolver signal when the acquisition azimuth coincides with the tracking azimuth, in the track mark generator channel, develops a V-section modulation in the output of V1A. This is the V-section signal which is desired. When the acquisition antenna has rotated 180° from that position, a second null occurs in the E-W resolver signal. This null also produces a V-section signal in the output of V1A, but this signal must be rejected. The voltage waveform necessary to select the desired signal and to

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reject the unwanted one is developed in the mixer amplifier stage. Two inputs are applied to this stage, the 4-kc carrier voltage and N-S resolver signal. The angle of difference between the 2 azimuth angles is zero when the two angles coincide. The cosine of this zero difference angle is +1, which would seem to indicate that at this instant the N-S resolver signal would be in phase with the 4-kc carrier. However, a 180° phase inversion which takes place in the azimuth channel causes the N-S resolver signal to be 180° out of phase with the 4-kc carrier at the instant when the azimuth angles of the two radars coincide. These two input signals are applied to opposite ends of 4-kc ADJ potentiometer R20, which is a mixing device. The brush arm of R20 is positioned so that zero voltage will be present at the instant when the N-S resolver signal is of maximum amplitude and is opposite in polarity to the 4-kc carrier. A single null will then occur during each revolution of the antenna. This null coincides with one of the two nulls which appear in the E-W resolver signal. When the undesired null in the E-W voltage occurs, the amplitude of the signal obtained at the mixing circuit is maximum. The relative time of occurrence of these nulls makes it possible to select the desired null in the E-W signal and to reject the undesired null which occurs 180° later. This single null occurs when the azimuth angles of the 2 radars are the same. Throughout the remainder of each revolution of the acquisition antenna, the mixed signal present at the brush arm of R20 will be a 4-kc signal in phase with the 4-kc carrier. The amplitude of this signal varies sinusoidally, being zero at the instant when the acquisition azimuth coincides with the tracking azimuth, and maximum when the acquisition azimuth is 180° different from the tracking azimuth. The mixed signal is coupled through capacitor C14 to the control grid of the tube. The d-c level at the control grid is -13 volts, as determined by voltage divider R13-R15. Because of the bias on the tube, the positive excursion of the signal developed across plate-load resistor R17 is clipped. This output signal is coupled through V13 for application to rectifiers in both the 40° detector and the 0.8° detector.

117. THE 40° DETECTOR

a. General. The 40° detector includes rectifiers V4A and V4B and an R-C filter network. The outputs of the 40° limiter and the mixer amplifier provide the 2 input signals. A single rectifier would be sufficient to detect the 40° V-section signals in the output of V1A. However, as 2 of these signals occur during each revolution of the acquisition antenna, a means of selecting the desired one and rejecting the undesired one is necessary. This requirement explains the purpose of the second input obtained from mixer amplifier V3A.

40° signal generated in the azimuth gate channel. The output of V1B is coupled through C2, developed across R11, and applied to the cathode of rectifier V2B in the 0.8° detector.

120. THE 0.8° DETECTOR

a. General. This detector is similar to the 40° detector. It consists of diode rectifiers V2A and V2B, together with an R-C filter circuit common to both diodes. One of the 2 input signals is the mixed signal obtained at the output of V3A. The other input signal is the modulated output of V1B, containing two 0.8° V-sections per revolution of the acquisition antenna.

b. Operation. The cathode of V2B is at a positive d-c level established by the setting of GATE ADJ potentiometer R10. This voltage may be adjusted within limits of zero and +44 volts. In order for V2B to conduct, the negative alternation of the signal applied to its cathode must be of sufficient amplitude to cause the cathode to be driven negative with respect to the plate. Since the d-c level at the cathode may be adjusted, it is possible to control the instant during the V-section signal when the decreasing amplitude of the signal will no longer cause conduction through the diode. This control operates to determine the effective width of the 0.8° signal which appears across the filter as a positive-going waveform. The effect of this control is to vary the position of the azimuth mark relative to the azimuth gate so as to cause the radial line of the electronic cross to appear in the visual center of the arc. Diode rectifier V2A operates in a manner similar to the V4A in the 40° detector, receiving the same input signal and producing effectively the same output signal. The 4-kc component of both input signals is filtered through capacitor C3, and the detected negative signal appearing across the filter is similar to that appearing at the output of the 40° limiter. The exception is that the 2 positive going pulses have a duration equivalent to only 0.8° of rotation of the acquisition antenna. One of these 0.8° pulses occurs at the instant when the negative voltage on which it rides is most positive (least negative). This signal coincides with the midpoint of the azimuth gate. The amplitude of this signal is sufficient to overcome the cutoff condition at the suppressor grid of the preknock selector, the stage to which it is applied. The other 0.8° signal occurs at the most negative point and this signal will not operate the preknock selector V7.

121. PREKNOCK SELECTOR V7

a. General. The azimuth mark appears once during each revolution of the acquisition antenna. This signal is produced by circuits of the mark generator channel which are triggered by a selected preknock pulse. It is the function of the preknock selector stage to select the preknock pulse which will be used as a trigger to produce the azimuth mark at the desired instant. V7 is

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azimuth is 180° from the tracking azimuth. This pulse occurs when the voltage level on which it is riding is most negative, and it will be rejected in the circuits to which it is applied.

118. AZIMUTH GATE CATHODE FOLLOWER V3B

The negative output of the 40° detector is applied directly to the control grid of V3B. This cathode follower serves to isolate the detector stage from the circuits to which the azimuth gate is applied. The plate is connected directly to the +150-volt supply and the large cathode resistor is connected to the -250-volt supply. The output signal obtained at the cathode is a faithful reproduction of the applied signal. The mean d-c level in the output is slightly less negative than the input and is approximately -21 volts. From the acquisition mark generator, the output of this stage (the acquisition azimuth gate) is applied to the precision indicators when the associated TRACK-ACQ switch is in the ACQ position. In the circuits of the precision indicators, only the most positive portion of this signal will operate the circuit to which it is applied. This portion is the equivalent of 30° of antenna rotation. This is also true of the tracking azimuth gate which is generated in the tracking mark generator channel and is applied to the designator presentation system circuits when the TRACK-ACQ switch is in the TRACK position. In addition, the tracking azimuth gate is also applied to the mixer channel. In that channel, only the most positive (least negative) 10° portion of the signal will operate the circuit. This signal aids in producing the 10° arc of the electronic cross.

119. THE 0.8° LIMITER V1B

a. General. The 40° waveform employed in the azimuth gate channel is initiated 20° before and ends 20° beyond the azimuth at which the azimuth mark must appear. To produce the azimuth mark at the desired instant, a signal must be developed which will occur at the midpoint of the 40° waveform. This signal is initially produced by the 0.8° limiter.

b. Operation. The modulated 4-kc output of V1A, containing two 40° V-sections per revolution of the acquisition antenna, is applied to V1B through the R-C coupling circuit consisting of C1 and R8. Cutoff limiting and grid limiting operate together to reduce the width of the V-section signal in the output to a value considerably less than 40° . The operating point of the tube is established by the cathode-biasing circuit, R6 and C15. The output signal is a modulated 4-kc signal of constant amplitude throughout most of the rotation of the acquisition antenna. At 2 points 180° apart the amplitude of the 4-kc signal decreases to zero value, reverses phase, and then returns to its initial value. One of these V-section signals occurs at the approximate center of the

b. Operation. The series of negative preknock pulses at the plate of the preknock selector is coupled to the grid of the conducting B-section through C11 and across R34. The first pulse having an amplitude greater than 10 volts will trigger the multivibrator. The A-section will then conduct, generating a negative-going signal across R36. Capacitor C10 has been charged during the quiescent period to the voltage difference existing between grid 7 and +250 volts. As this charge cannot change instantaneously because of resistance in series with C10, the negative-going signal at plate 1 is coupled to grid 7, cutting off the B-section. The plate of the B-section then goes to a potential of +250 volts, where it remains until capacitor C10 can discharge sufficiently to return the grid of the B-section to a potential above cutoff. The time constant of the discharge of C10 is such that the B-section remains cut off for 20,000 microseconds, and a positive square wave of that duration is developed at the plate of the B-section once during each revolution of the acquisition antenna. The leading edge of this signal coincides with the midpoint of the azimuth gate. Negative preknock pulses which are applied to grid 7 after the multivibrator has been triggered have no effect as the grid of the B-section is below cutoff at that time.

123. AZIMUTH MARK MULTIVIBRATOR V5

a. General. This stage is a one-shot multivibrator which generates a positive 612-microsecond square wave. This signal is the azimuth mark. The positive 20,000-microsecond square wave input is differentiated in an R-C peaking circuit and the positive peak occurring at the leading edge of the input square wave triggers the multivibrator. The acquisition azimuth mark, developed in the acquisition mark generator channel, is applied to the mixer channel. It appears on the acquisition indicators as the flashing steerable azimuth line. The tracking azimuth mark, developed in the tracking mark generator channel, is also applied to the mixer channel. This signal, together with the tracking range gate, acts to produce the radial line of the electronic cross.

b. Operation. The positive 20,000-microsecond square wave generated by V6 is applied to the grid of V5A through the R-C peaking circuit consisting principally of resistor R29 and capacitor C9. Differentiation develops a positive peak which is applied to the grid of V5A coincident with the leading edge of the applied 20,000-microsecond square wave. Prior to the appearance of this signal, the A-section is held below cutoff by a negative 31-volt potential obtained at the junction of voltage-divider resistors R24 and R25. The B-section is conducting heavily, with its cathode at ground and its control grid returned to the +250-volt supply through resistor R30 and MARK LENGTH control R32. The positive peak resulting from the differentiation of the leading edge of the applied square wave overcomes the cutoff potential at the control

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a coincidence tube. The preknock pulse is applied as one input. The output of the 0.8° detector provides the other input. This latter input overcomes the cutoff condition of the tube at the time of the desired 0.8° positive-going signal. During the conduction period of V7, the preknock pulses appearing on the grid are amplified and inverted. The output, which consists of a series of negative preknock pulses, provides the input to V6.

b. Operation. The impedance of the cable from the pulse synchronizer is matched by R41. The preknock pulse is coupled to the control grid of the tube through C12. The signal is developed across parallel resistors R39 and R40, which form a voltage divider between the -250-volt supply and ground. This voltage divider develops -15 volts bias for V7. Although each preknock pulse overcomes the control-grid cutoff, plate current may flow only when the suppressor grid rises to a less negative potential. This occurs for 0.8° of antenna rotation at the midpoint of the azimuth gate. If the antenna rotates at 20 rpm, the width of the 0.8° signal at the suppressor grid will be over 6,500 microseconds. During this period, 6 or 7 preknock pulses will appear at the control grid and will be amplified and inverted by the tube. The output of V7, therefore, consists of a series of negative preknock pulses. This series occurs once per revolution of the acquisition antenna at the midpoint of the azimuth gate.

122. THE 20,000 MICROSECOND MULTIVIBRATOR V6

a. General. With the acquisition antenna rotating at 20 rpm, the 0.8° signal has a duration of over 6,500 microseconds. During the time this signal is applied to V7, 6 or 7 preknock pulses occur. All of these preknock pulses appear in the output of that stage. If this output were applied directly to the azimuth mark multivibrator, which generates a 612-microsecond signal, as many as 8 consecutive azimuth marks could be generated. Tube V6 acts to prevent such successive triggering of the azimuth mark multivibrator. The output of the preknock selector is applied to V6. The first in the chain of preknock pulses to exceed 10 volts in amplitude serves to trigger V6, which then produces a positive rectangular waveform having a duration of 20,000 microseconds. The duration of this signal is sufficient to insure that the return of V6 to the quiescent state can occur only after the disappearance of the input signal obtained from the preknock selector. Multiple generation of the azimuth mark is thereby prevented. The positive 20,000-microsecond square wave is applied to the azimuth mark multivibrator V5. This stage V6 is a one-shot, cathode-coupled multivibrator. The grid of the B-section (pin 7) is returned to a positive d-c level of 32 volts. This potential is obtained from the voltage divider R31-R33. As a result, the B-section conducts in the quiescent state. Plate current in the B-section develops a positive potential at the cathode. With the grid of the A-section grounded, this positive potential is sufficient to bias the A-section approximately 10 volts below cutoff.

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of the range circle and the steerable azimuth line. The precision indicators locate target signals with respect to the azimuth and range marks more accurately than is possible on the PPI's. In addition to their primary function, the precision indicators make it possible to distinguish targets separated in range by 200 yards or in azimuth by 35 mils. The precision indicators utilize signals from every part of the target designate system in order to generate the desired sweep. As a result, the precision indicators make good troubleshooting devices since a failure in almost any component of the system will show up here. The range and azimuth sweep channels are responsible for generating the B-type scan on the 5-inch PI tube. The video amplifier allows this sweep to be intensity modulated with video and marks. The unblanking channel causes the PI to be illuminated at the proper time.

126. RANGE SWEEP GENERATOR CHANNEL (TM 9-5000-26 pages 233 and 236)

a. General. The B-type presentation seen on the precision indicators is produced by a combination of range and azimuth sweep signals. Vertical deflection (bottom-to-top) of the electron beam within the cathode ray tube produces range sweeps of 5,000 yards duration (approximately 30 microseconds). These range sweeps are generated at the pulse recurrence frequency of the equipment but serve no purpose except during the 30° period when the tube is unblanked. During that period, application of azimuth sweep signals to the deflection-coil assembly tends to deflect the electron beam across the face of the tube from left to right. This action is relatively slow, producing a single full deflection to the extreme right edge of the scope in the time required for the acquisition antenna to rotate through 30°. At a rotational rate of 30 rpm, a total of 170 range sweeps will be generated. The effect of this combination of sweep signals is to cause each range sweep to occur to the right of the preceding sweep. The first of these 170 range sweeps occurs at the extreme left of the scope. Each succeeding sweep occurs farther to the right with the last of the 170 range sweeps occurring at the extreme right edge of the scope. The 5,000-yard vertical range sweeps are produced as a result of sweep signals generated in the range sweep channel. The tracking range gate or the acquisition range gate is applied to trapezoidal sweep generator V1A. This signal cuts off V1A causing the generation of a positive trapezoidal voltage waveform. This signal is then amplified by two stages, V2 and V3. The positive output signal of V3 is applied to the output amplifier stage which consists of tubes V4 and V5 in parallel. The control grids of the output amplifier are clamped at a fixed negative potential to insure that each range sweep will start from a uniform level. The plate load of the output amplifier stage is the vertical deflection winding of the precision indicator. A linear range sweep is obtained

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grid of the A-section. Current flow through the A-section develops (at plate 1) a negative-going signal which is coupled through C8 to the grid of the conducting B-section. This reduces the current through the B-section, causing a rise in plate voltage. This rise is coupled through C7 to the grid of the A-section, further increasing the current flow through the A-section. This regenerative process will cut off V5B almost immediately and at the same time will drive V5A into heavy conduction. The A-section will continue to conduct for a period of 735 microseconds, as determined by the adjustment of MARK LENGTH control R32 which is adjusted so that the B-section is held below cutoff for 612 microseconds. By means of this adjustment, the length of the flashing steerable azimuth line may be controlled. During this period, the conduction of the A-section will develop a positive square wave across its cathode resistor, R27. Obtaining the output signal across this 1,000-ohm resistor allows the cable impedance to be matched without the addition of a cathode follower stage.

124. FIELD ADJUSTMENTS

a. 4-kc adjustment. With the test amplifier connected to TPI on the mark generator chassis, adjust the 4-kc control until one complete null in signal occurs per revolution of the antenna.

b. Gate adjustment. With the precision indicator in the track position, adjust the track mark generator GATE ADJ until the radial line of the electronic cross is centered on the arc. This control has no effect on the acquisition mark generator.

c. Mark length adjustment. With the PPI's set for 120,000 yards and the track range set at 99,900 yards, set the acquisition mark generator MARK LENGTH control so that the flashing steerable azimuth line extends to 120,000 yards and set the MARK LENGTH control on the track mark generator such that the radial line of the cross is completely visible.

Section IV. DESIGNATE PRESENTATION SYSTEM

125. INTRODUCTION (TM 9-5000-26, page 233)

The precision indicators provide the link between the acquisition and tracking radars. This link between the two radar systems makes it possible for the battery control officer to designate selected targets accurately and for the tracking radar to get on target quickly. The precision indicators accomplish this by presenting an expanded display of the marks and video signals appearing in a selected 30° by 5,000-yard sector at the target tracking console and 30° by 8,000-yard sector at the battery control console. This sector may be centered either at the intersection of the electronic cross or at the intersection

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d. Clamper V1B. Tube V1B is a cathode follower which operates to maintain a constant d-c reference level at the control grids of the power amplifier during the no-signal period. This is necessary in order to cause the electron beam to start from the same level on each succeeding range sweep. The grid of V1B is held at approximately -83 volts by voltage divider R12-R13. The cathode potential, which is determined by current flow through V1B and cathode resistor R11, is slightly less negative than the control grid being about -70 volts. The -83-volt, control-grid potential is held constant by the action of capacitor C5B. The positive-going leading edge of the trapezoidal voltage waveform at the plate of V3 is coupled through capacitor C4 to the cathode of V1B. The amplitude of this voltage change is sufficient to cut off V1B. The positive trapezoidal voltage waveform appearing at the control grids of V4 and V5 is developed as a result of charging current flowing through R11 to C4.

e. Power amplifier V4 and V5. The power amplifier employs two 6AR6 tubes in parallel in order to deliver the desired amount of current to the deflection coil which forms the plate load. Resistors R18, R19, and R24 prevent the parallel tubes from breaking into undesired oscillations. A plate-supply voltage of +450 volts is used, with a screen-supply voltage of +320 volts. These comparatively high-supply potentials are for the purpose of obtaining a large value of current flow during the conduction period. During the no-signal period, the tubes are cut off by the negative 70-volt potential present at the cathode of clamper V1B. The positive trapezoidal voltage coupled from the plate of V3 through C4 appears at the control grids of both tubes. The applied voltage waveform is such as to cause the tubes to deliver a linear sawtooth current through the vertical deflection coil of the cathode ray tube. The coil is shunted by resistor R7 to prevent oscillations in the coil. To insure linearity of current flow through the deflection coil, distortion introduced by the various stages of the range sweep channel must be eliminated. This is done by degenerative current feedback from the power amplifier stage to V2, the first sweep amplifier tube. Resistors R23 and R27 in parallel provide the cathode load for tubes V4 and V5. These same resistors, together with resistor R17 and capacitor C6, are in the cathode circuit of sweep amplifier V2. The positive signal at the grid of V2 produces a positive signal at the grids of V4 and V5. As R23 and R27 are unbypassed, the cathode potential of V4 and V5 rises with the grid. As a result, the cathode of V2 also rises. Therefore, the grid-to-cathode signal at V2 is equal to the grid-to-ground signal minus the feedback signal from V4 and V5. The feedback improves the linearity of the current waveform applied to the deflection coil by overcoming the distortion which would otherwise be introduced by V2, V3, V4, and V5. A component of current through R23 and R27 flows through the screen-grid circuit and not through the deflection coil. To prevent this component of current from producing a corresponding feedback signal, a feedback voltage is

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by applying a linear sawtooth current to the vertical deflection coil. To improve the linearity of this current waveform degenerative feedback is employed between the output amplifier and amplifier V2.

b. Trapezoidal sweep generator V1A. Tube V1A is a triode operating as a diode. The tube conducts in the quiescent state. The very large plate resistance causes the voltage at the plate of the tube to be only slightly above ground potential. As a result, capacitor C1 has only a very small charge. The positive 30-microsecond range gate is applied to the cathode of V1A at the pulse recurrence frequency of the equipment. The range gate cuts off the tube and capacitor C1 begins to charge toward the supply potential of +150 volts. The time constant of the charge path of C1 is approximately 4,700 microseconds. However, C1 is allowed to charge only for the 30-microsecond period during which the tube remains cut off. This period of charge is less than 1 percent of the time constant, and the sawtooth component of the waveform appearing at the plate of V1A will therefore be linear. Capacitor C1 is not returned to ground but to the junction of resistors R3 and R4 in the cathode circuit. The rectangular gate applied to the cathode of the tube will appear at greatly reduced amplitude at the junction of R3 and R4. This rectangular component is coupled through C1 and appears at the plate of V1A in conjunction with the sawtooth voltage which results from the charging of C1. These two components together produce a trapezoidal voltage waveform at the plate of V1A. This signal is directly coupled to the grid of the following stage. The rectangular component of this signal is required to overcome the inductance of the vertical deflection coil to which the range sweep signal is applied. At the termination of the range gate signal at the cathode, V1A resumes conduction and quickly discharges C1 to the quiescent state level. The extent of the sweep deflection which will appear on the scope is determined by the amplitude of the sweep signal. The AMPLITUDE control, R2, provides a means of changing the charging time constant of C1 and hence determines the amplitude of the trapezoidal voltage generated by V1A. The AMPLITUDE control is adjusted to obtain a range sweep which extends from the bottom to the top of the aperture of the cathode ray tube.

c. Sweep amplifiers V2 and V3. These 2 stages are identical to stages V1 and V2 of the PPI sweep amplifier chassis and are described in detail in the special text on the PPI. The positive trapezoidal voltage applied to V2 appears again at the output of V3 with the addition of a positive spike at the leading edge of the waveform. This output is coupled through C4, is clamped at a negative reference level by the action of clamper V1B, and is applied to the power amplifier stage, V4 and V5.

signals are applied to a ringdemodulator consisting of diodes V1A, V1B, V2A, and V2B. This circuit, together with two R-C filters, detects the modulation of the E-W resolver signal. This modulation is normally sinusoidal, with 1 cycle of modulation being completed during each revolution of the acquisition antenna. Two low-frequency sine waves 180° out of phase are produced. Most of the positive alternations of these 2 signals are clipped by limiters V3A and V3B and the resultant flat-top waveforms are applied to the grids of push-pull amplifiers V4 and V5. The applied signals cut off V4 and V5 at such an instant that current flow through the tubes changes only during 30° of antenna rotation. This push-pull current change is applied to the horizontal deflection windings of the precision indicator tube and causes deflection of the sweep to the right or to the left. The change of current flow through the push-pull amplifiers and the horizontal deflection windings happens twice during each revolution of the acquisition antenna each time the E-W resolver signal experiences a null. The azimuth mark (about which the precision indicator display is centered in azimuth) is generated as a result of the appearance of 1 of these 2 nulls. The current change occurring at that time produces the desired left to right deflection of the sweep. After 180° of antenna rotation, a push-pull current change in the opposite direction occurs. This current change would tend to deflect the sweep from right to left but this effect is not visible as the precision indicator is then cut off.

b. Ring demodulator V1 and V2. The operation of this stage is similar to the operation of the ring demodulators of the N-S and E-W sweep channels of the acquisition presentation system. The ring demodulator receives 4-kc carrier reference voltage and the E-W resolver signal. This latter signal is obtained either from the track azimuth resolver or from the azimuth line resolvers depending upon the position of the TRACK-ACQ switch associated with the precision indicator. The demodulator detects the modulation of the E-W resolver signal, producing 2 output sine waves of opposite phase. The period of these sine waves is equal to a single rotation of the acquisition antenna. In the demodulators of the acquisition presentation system, the detected sine waves were developed across resistors which were returned to ground, and hence the detected sine waves varied about ground potential. However, in the ring demodulator of the precision azimuth sweep generator, these sine waves are developed across resistors R8 and R9 which are returned to a negative potential of approximately 9 volts. This -9-volt potential is obtained from the junction of resistors R7 and R11 in the voltage divider, R7-R11-R13, and is held constant by capacitor C1. A negative potential of approximately 2 volts is present at the junction of R11 and R13. Resistors R10 and R14 with capacitors C2 and C5 form filter circuits to remove the 8-kc ripple present in the output of the ring demodulator. A similar filter consisting of R12 and R15 and C3 and C4 is also used. A sine wave riding a negative d-c level of 9 volts is present at the junction of R10 and R14. A sine wave which also rides at a

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obtained from the screen-grid circuit which is opposite in phase to the feedback signal developed across the 3 cathode resistors. The flow of screen-grid current through resistor R25 produces a negative signal which is fed back through resistor R20. The amplitude of this negative signal is sufficient to cancel the positive feedback component developed across the cathode resistors as a result of screen-grid current. Capacitor C6 couples the steep initial voltage rise at the cathode of V4 and V5 to the cathode of V2. As a result, a spike of very short duration appears in the voltage waveform at the instant of the initial transient. It is this voltage spike which overcomes the inductance of the deflection coil. Resistor R26 and capacitor C5C form a decoupling circuit between the screen-grid circuit of the power amplifier and the positive 320-volt supply.

127. AZIMUTH SWEEP GENERATOR CHANNEL (TM 9-5000-26, page 237)

a. General. The azimuth sweep signal required to move the range sweep across the face of the scope from left to right is developed in the azimuth sweep generator. This unit (fig 32) receives 2 input signals, one of which is the 4-kc carrier reference voltage. The second input is the E-W output signal of either the track azimuth resolver or the azimuth line resolvers. These two 4-kc input

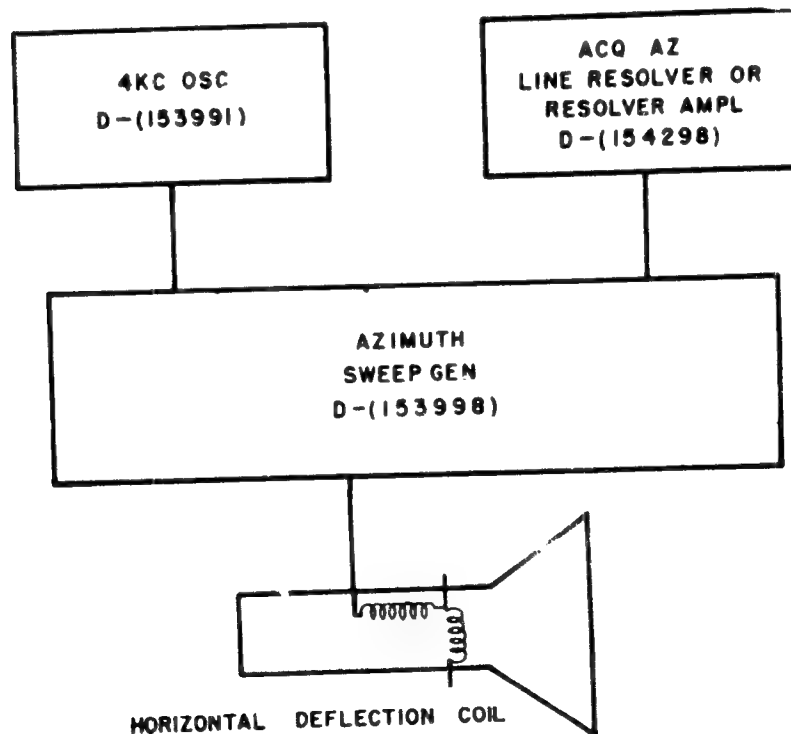


Figure 32. Azimuth sweep generator, block diagram.

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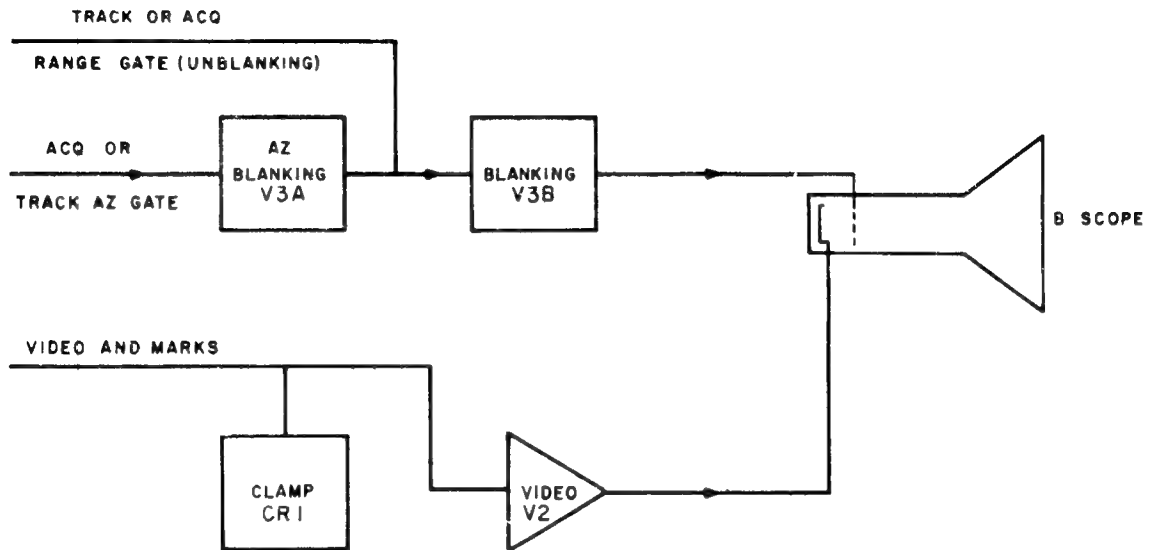


Figure 33. Precision video amplifier.

128. VIDEO AMPLIFIER

a. General. The video amplifier chassis contains an unblanking channel (fig 33) and a video amplifier stage. The unblanking channel includes azimuth gate amplifier V3A and unblanking mixer V3B. Except during the presentation of the comparatively small sector which is displayed, the precision indicator tube is held at visual cutoff. The tube is unblanked only for a 5,000-yard portion of those sweeps which occur during a selected 30° arc of antenna rotation. The purpose of the unblanking channel is the production of a positive unblanking signal for application to the control grid of the precision indicator tube. Depending upon the position of the TRACK-ACQ switch, either the tracking range and azimuth gates or the acquisition range and azimuth gates are applied to the unblanking channel. The range gate, occurring once during each repetition period is applied to unblanking amplifier V3B. However, this signal alone is not sufficient to cause V3B to conduct. The azimuth gate is applied to the cathode of azimuth gate amplifier V3A once during each revolution of the acquisition antenna. The positive 30° output signal of V3A provides a second input to unblanking amplifier V3B. During the period of the 30° gate, the 5,000-yard range gates applied to V3B appear at the plate as positive rectangular waveforms.

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negative d-c level of 9 volts but is of opposite phase is present at the junction of R12 and R15. These signals are applied to the plate circuits of the two sections of limiter V3.

c. Limiters V3A and V3B. The two sections of V3 operate in identical fashion so only 1 section will be discussed. The cathode of the diode is at a potential of -2 volts obtained from voltage divider R7-R11-R13. During a part of the positive alternation of the applied sine wave, the signal at the plate of the diode tends to become more positive than the -2-volt cathode potential. At that time, current flow through the diode clips the waveform. The signal appearing at the plate of the diode, therefore, is a sine wave with most of the positive alternation clipped.

d. Push-pull amplifier V4 and V5. The output signals of limiters V3A and V3B are applied to the control grids of V4 and V5. Both these tubes are dual-triodes with the 2 triode sections of each tube operating in parallel to provide the necessary amount of current to the deflector-coil windings which form their plate loads. Resistors R18 through R25 are designed to suppress parasitic oscillations. The stages will be cut off whenever the grid potential becomes more negative than -16 volts. During the time that V4 is conducting (with a potential of -2 volts present at its grid), V5 is cut off. When V5 is conducting, V4 is cut off. In both these conditions, the current flow through the two tubes is not changing and no change occurs in the magnetic field about the horizontal deflection winding. As the grid potential at one tube goes from -2 volts to cutoff, the current through that tube decreases to zero. At the same time, the grid potential at the other tube rises from cutoff to -2 volts and current through that tube rises from zero to maximum. As this push-pull current change occurs, the net magnetic field about the horizontal deflection coil changes direction. As a result, the sweep of the cathode ray tube is deflected from one side of the scope to the other. After 180° of antenna rotation, changes in the opposite direction occur at the control grids of V4 and V5, and the sweep is deflected back in the opposite direction. The desired deflection from left to right occurs at the sector in which the cathode ray tube is unblanked. The undesired right-to-left deflection occurs when the cathode ray tube is cut off and hence is not visible on the scope. The BAL ADJ control R17 in the cathode circuit of V4 and V5 provides a means of obtaining equal current flow through these two tubes. In addition, it is necessary that the change in grid voltage of V4 and V5 from -2 volts to cutoff (and vice versa) occurs during a period equal to 30° of antenna rotation. This may be established by properly controlling the amplitude of the signals applied to V3. This is the function of R2, the AZ ADJ control, in the ring demodulator circuit.

unblanking amplifier V3B. The cathode of V3B is connected in the voltage divider at the junction of R5 and R7, where a -4-volt potential exists. The voltage at the plate of V3A and the control grid of V3B is sufficiently negative to keep V3B cut off during the period between the azimuth gates. The positive range gate is applied through capacitor C7 to the cathode of V3B each repetition period. While V3B is held cut off by the conduction of V3A, application of the range gate has no effect upon the tube; and the plate voltage remains at the value of the supply voltage, or +150 volts. Since the cathode of V3B is connected in the voltage divider, application of the range gates to the cathode causes instantaneous voltage changes to occur along the entire voltage divider. To prevent these variations from affecting the circuit action, capacitors C2 and C3 are connected to ground from the grids of both sections of V3.

d. Operation during unblanked period. Azimuth gate amplifier V3A is cut off once during each revolution of the acquisition antenna when the signal applied to its cathode becomes least negative. This signal becomes least negative during the azimuth gate when the positive-going 40° waveform superimposed on the varying d-c potential coincides with the least negative value of the varying potential. When V3A is cut off, the potential at its plate rises to approximately -7 volts. This same change occurs at the control grid of unblanking amplifier V3B, overcoming the cutoff bias normally applied to that tube. The instant when this cutoff bias is overcome is controlled by the adjustment of AZIMUTH BLANK ADJ R2, which is set so that V3B is permitted to conduct for a period equal to 30° of antenna rotation. When V3B conducts, the plate voltage decreases from +150 volts to some lower potential. The time required for the acquisition antenna to turn through 30°, at a rate of rotation of 30 rpm, is such that 170 range gates are generated and applied to the cathode of V3B, causing the plate potential to return to +150 volts each time a range gate is applied. During each revolution of the acquisition antenna, signals appear at the plate of V3B only during the 30° of antenna rotation which correspond to the appearance of the azimuth gate. During this 30° period, a series of positive 30-microsecond square waves appear at intervals of 1,000 microseconds. At the plate of V3B, the maximum positive potential attained by these square waves is +150 volts.

e. Control-grid circuit of the cathode ray tube. In the absence of any signals at the plate of V3B, the precision indicator cathode ray tube is cut off. A positive potential is present at its cathode and the control-grid potential is sufficiently less positive to insure that the tube will not conduct. This control-grid potential is obtained at the brush arm of INTENSITY control R5, which is connected in a voltage divider between +150 volts and -250 volts. The control-grid potential may be varied between the limits of +60 volts and -30 volts. The

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These signals are coupled to the control grid of the precision indicator, bringing the tube almost to the cutoff level. The video amplifier stage V2/C10 receives the positive output of the video and mark channel. After amplification and inversion, these signals are applied directly to the cathode of the precision indicator tube. Application of negative video or marks to the cathode of the tube when an unblanking signal is present at the control grid will cause the cutoff condition to be overcome. As a result, the signals will appear on the screen of the tube as intensified portions of the sweep.

b. Video-amplifier stage V2. The positive output signals of the video and mark channel are applied through GAIN potentiometer R9 (located on the front panel of the precision indicator) to the control grid of video amplifier V2. The precision video amplifier is identical to the PPI video amplifier stage. For a complete discussion of the operation of the stage, see the PPI special text. The stage amplifies and inverts the input signals. The plate of V2 is connected directly to the cathode of the precision indicator cathode ray tube. These negative signals, appearing during the presence of gating signals at the control grid, will overcome the cutoff condition of the tube. The applied video and mark signals will then appear as intensified marks on the face of the scope.

c. Unblanking channel during visual cutoff. Because the 2 stages V3A and V3B of the unblanking channel operate together, it is desirable to consider them together. The unblanking channel contains a voltage divider connected between -250 volts and ground. This voltage divider consists of resistors R1, R2, R3, R5, and R7. In the absence of any current flow through either section of V3, a negative potential of approximately 7 volts is present at the junction of R3 and R5. This potential is the plate-supply voltage for the azimuth gate amplifier, V3A. The control grid voltage for V3A is obtained from the brush arm of AZIMUTH BLANK ADJ control R2, and may be varied between -24 volts and -80 volts. The azimuth gate is applied to the cathode of V3A. This input is a negative d-c voltage which varies sinusoidally as the acquisition antenna rotates, completing 1 cycle during each revolution of the antenna. Superimposed on this negative voltage are 2 positive-going waveforms, each having a duration equivalent to 40° of antenna rotation. One of these superimposed waveforms appears when the varying negative voltage is least negative (most positive), and this waveform is termed the azimuth gate. The mean d-c level of the signal, except during the azimuth gate, is considerably more negative than the -7-volt, plate-supply potential and hence V3A conducts throughout most of each revolution of the antenna. The plate-load resistor, R4, is quite large. Because of its large value, very little voltage is dropped across V3A and practically all the signal appears across R4. The current through V3A, flowing through R4, causes the voltage at the plate of V3A to be considerably more negative than the -7-volt supply potential. The plate of V3A is connected directly to the control grid of

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recurring 30-microsecond signal which appears at the plate of V3B during the azimuth gate is applied to the R-C coupling circuit consisting of capacitor C4 and resistor R12. The positive 30-microsecond range gates, superimposed on this d-c level, raise the potential at the control grid of the cathode ray tube. The INTENSITY control is adjusted so that the potential attained by the range gates will be almost sufficient to overcome the cutoff condition of the tube. The appearance of negative video and mark signals at the cathode of the tube will then cause conduction.

129. PRECISION INDICATOR TUBE (TM 9-5000-26, page 233)

The precision indicator tubes are 5FP7A five-inch cathode ray tubes in which a magnetic deflection system is employed. The tube is cut off in the absence of an unblanking signal. The d-c level present on the control grid is determined by the setting of the INTENSITY control. This potential may be varied within the limits of -30 volts and +60 volts. The control is adjusted so that the application of unblanking signals to the control grid will almost overcome the visual cutoff condition. Mixed video and marks are applied as negative signals to the cathode, reducing the positive potential at the cathode an amount sufficient to overcome the cutoff condition. The tube then conducts, and the video and marks appear as illuminated marks on the face of the scope. Acceleration of the electron beam within the tube is accomplished by a positive 320-volt potential applied to the first anode and a positive 5,000-volt potential applied to the aquadag coating. This high voltage is obtained from the indicator high-voltage power supply discussed in the special text on the PPI. Focusing of the electron beam is done by adjusting the physical position of a permanent magnet. The desired sweep is obtained by application of azimuth sweep signals to the horizontal deflection windings and range sweep signals to the vertical deflection coil. To insure that the range sweep will consistently start from a uniform level at the bottom edge of the display, a second permanent magnet is utilized. This is the decentering magnet. It is adjusted to cause the electron beam to return to the bottom of the display following each range sweep. Adjustments applicable to this circuitry can be found in TM 9-5000-26 and BTL, vol VI, sec 12-VIII-5.

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